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Transceiver and System Design for Digital Communications

5th Edition

Scott R. Bullock

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To my loving wife, Debi;
to Crystal, Cindy, Brian, Andy, and Jenny;
and to my mother, Elaine.

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Preface

This book covers basic communications theory and practical implementation of transmitters and receivers. In so doing, the focus of this book is on digital modulation, demodulation methods, probabilities, detection of digital signals, spread-spectrum system design and analysis, radar and radar communications, volume search, acquisition, track and cognitive system processes. This book was written for those who want a good understanding of these basic principles. It also provides an intuitive and practical approach to digital communications with a hands-on approach including several examples for the reader to really understand these wireless techniques. Therefore, it is a valuable resource for anyone involved in wireless communications, networks, radar and cognitive systems. The reader will gain a broad understanding of these topics with examples of many types of commercial and military data link systems.

Chapter 1 describes transceiver design using a link budget to analyze possible trade-offs. This includes tracking of signal power and noise levels in the system, calculation of the noise figure, the gains and losses of the link, and the required signal level with respect to noise, including the link margin to provide the specified probability of error. This chapter also discusses frequency band designations along with the definitions and uses of decibels. Doppler and how it affects the link margin is covered. Spread-spectrum techniques and process gain, as well as coding gain, are discussed. The chapter concludes with an example of a link budget and a transceiver design that coincides with it.

Chapter 2 evaluates the basic functions of the transmitter, including antennas, transmit/receive (T/R) control, classes of power amplifiers, Crest factor, the upconversion process, sum and difference frequencies and the requirement to eliminate one of the conversion products, voltage standing wave ratio, and maximum power transfer principle. This chapter discusses advantages of digital versus analog communications. This chapter focuses on digital modulation techniques including phase-shift keying (PSK) and frequency-shift keying, phasor constellations and noise immunity, and error vector magnitude as a quality metric of the transmission. This chapter also addresses the advantages of continuous PSK modulation and spectral regrowth, saturation, and spectral efficiency. Shaping filters for the digital waveforms are discussed, with ideal and practical solutions. Direct sequence spread-spectrum systems are addressed, along with the advantages of using spread spectrum, including process gain and antijam with jamming margin. Maximal length sequence codes, including Gold codes and others, are included, along with spectral lines that are generated in the frequency domain and how to

remove them. Several digital modulation techniques using PSK are provided, along with block diagrams and phasor diagrams to help analyze the different types of PSK systems that are used today. Variations of PSK systems and other types of spread-spectrum systems are also discussed, such as frequency hopping, time hopping, and chirped frequency modulation. In addition, multiuser techniques are explained, including time, code, and frequency access systems. Finally, orthogonal techniques including orthogonal frequency division multiplexing are considered, along with power control to reduce near–far problems.

Chapter 3 covers the basic functions of the receiver, including the antenna, T/R control, image-reject filter, low-noise amplifier, downconversion process, third-order intercept, image frequency, and various methods to determine dynamic range. Phase noise, mixers, spur analysis, bandwidths, and filters are also addressed. The discussion of digital processing includes principles such as group delay, sampling theorem and aliasing, antialiasing filters, and analog-to-digital converters (ADCs) including piecewise linear ADCs.

Chapter 4 discusses the design and analysis of automatic gain control (AGC). The main elements of a good AGC design are provided, including the amplifier curve, linearizer, detector, loop filter, threshold level, and integrator for zero steady-state error for a step response. Control theory is used to define the stability characteristics of the AGC to design the optimal AGC for the system. Chapter 4 also details the phase-locked loop (PLL), particularly for the lock condition, and compares the similarities between the AGC and the PLL. Feedback systems and oscillations, including the Barkhausen criteria, are reviewed, along with Bode diagrams to determine both gain and phase margin.

Chapter 5 describes the demodulation process portion of the receiver, which includes the different methods of carrier recovery loops which includes the squaring loop, Costas loop, Modified or Hard Limited Costas loop, and Decision Directed Costas Loop, also includes detecting the CP-PSK modulation. This chapter describes how to correlate and detect the data using either a matched filter correlator or sliding correlator and maintaining lock of the code tracking loops using an early-late gate analysis. Pulse-position modulation and demodulation are discussed and also the advantages and disadvantages of coherent versus differential demodulation techniques. The chapter also discusses symbol synchronizer, the eye pattern, intersymbol interference, scrambler–descrambler, Shannon's limit for information channel capacity, and phase-shift detection for intercept receivers.

Chapter 6 contains a basic discussion of the principles of digital communications. This includes an intuitive and analytical approach to understanding probability theory, which is used in designing and analyzing digital communications. The explanation shows the basic Gaussian distribution and how to apply it to probability of error. Quantization and sampling errors, as well as the probability of error for different types of spread-spectrum systems, along with the curves and how to apply them in a design, are evaluated for the system. Also examined is the methods for detecting errors, including parity, checksum, and cyclic redundancy check. Error correction using forward error correction is assessed, including topics

such as interleaving, block codes, convolutional codes, systematic linear block codes, generation of these codes, and hamming codes. Also includes the probability of missed errors. The Viterbi algorithm, multi-hop, low-density parity check codes, and turbo codes are also discussed. In addition, this chapter provides basic theory on pulsed systems, which includes spectral plots of the different pulse types.

Chapter 7 focuses on multipath. This chapter discusses the basic types of multipath, including specular reflection of both smooth and rough surfaces and diffuse reflections off a glistening surface as well as the Rayleigh criteria for determining if the reflections are specular or diffuse. The curvature of the earth is included for systems such as those used for satellite communications (SATCOMs). The advantages of using leading edge tracking for radars to mitigate most of the multipath present are discussed. Several approaches to the analysis of multipath are provided, including vector analysis and power summation. Included are discussions on several different multipath mitigation techniques including using antenna diversity.

Chapter 8 describes several types of jammers such as spot, barrage, CW, cognitive and burst jammers including capturing the AGC and methods that improve the system operation against jamming signals. Antijam techniques are discussed including; burst clamps to minimize the effects of burst jammers, adaptive filtering to reject narrowband signals and a Gram-Schmidt Orthogonalizer (GSO) that uses two antennas to suppress the jamming signal. An in-depth analysis is provided on the adaptive filter method using the adaptive filter configured as an adaptive line enhancer using the least mean square algorithm. A discussion of the suppression results due to amplitude and phase variations is included as well. An actual wideband system, providing simulation and hardware results of the adaptive filter, are discussed. In addition, many different types of intercept receivers for detection of signals in the spectrum are included.

Chapter 9 covers cognitive and adaptive techniques to make a "Smart" cognitive system. This shows the various ways to monitor the environment and then to adapt to the changing environment to provide the best solution using the available capabilities. The basic cognitive techniques covered are dynamic spectrum access, which changes frequencies; adaptive power and gain control; techniques using modulation waveforms, spread spectrum, adaptive error correction, and adaptive filters; dynamic antenna techniques using active electronically scanned arrays (AESAs) for beam forming, null steering, adaptive beam spoiling and narrowing, and pointing. It also includes multiple-in and multiple-out capabilities, antenna diversity, and multipath communications. This chapter also discusses adapting networks using multihop for meshed network like mobile ad hoc networks, serial versus parallel evolution, total cognitive system processes for optimal solutions, flow diagrams incorporating reasoning and learning, and determining and selecting system capabilities. This also covers prediction capabilities, reasoning and learning capabilities, methods and elements of reasoning, multiagent systems, cooperative and noncooperative game theory, coalitional game theory, Nash equilibrium, and the challenges for implementing, governing, rules and regulations, and enforcement of cognitive systems.

Chapter 10 addresses directional volume search, acquisition, and track. There are several ways to perform volume search, this chapter address many of the techniques and performs the analysis and results for the optimal solutions for given requirements. It also discusses ways to improve the directional volume search using beam spreading and sidelobe detection. Additional requirements include coordinate conversion, directional beam and angular resolution. Acquisition is discussed using two-dimensional sequential scanning to acquire the target or user. Other methods include Conical scan and Monopulse for tracking targets and the performance comparison between them. Both sequential lobing when the target is in the beamwidth along with the alpha-beta tracker provides excellent performance for tracking the target once it has been acquired. Integration of multiple types of trackers especially for integration of closed-loop and open-loop tracking during coasting is addressed and can be weighted depending on the accuracy of the tracker.

Chapter 11 covers broadband communications and networking, including high-speed data, voice, and video. Many generations of mobile wireless communication products have evolved through the years and are designated as 1G through 4G with 5G on the horizon. Broadband is also used in the home to connect to the outside world without having to run new wires using power line communications, phone-line networking alliance, and radio frequency such as IEEE 802.xx and Bluetooth. Along with the distribution of information, networking plays an important role in the connection and interaction of different devices in the home. Worldwide Interoperability for Microwave Access is another radio frequency wireless link based on the IEEE 802.16 standard, and LTE has emerged as a viable solution for the next generation of wireless products. The military is investigating in several networking techniques to allow multiple users for communications, command, control, and weapon systems, including the Joint Tactical Radio System and Link 16. Software-defined radio; cognitive techniques; software communications architecture; five clusters development; the network challenge including gateways; and different network topologies including Star, Bus, Ring, and Mesh are all addressed.

Chapter 12 covers SATCOM for various applications in both the commercial and military sectors. The infrastructure for distributing signals covers the widest range of methods for communications. The most remote places on the earth have the means for communication via satellite. SATCOM has the needed infrastructure, coverage, and bandwidth for communication and networking. This can be combined with other types of communication systems to make this an ideal candidate for providing ubiquitous communications to everyone worldwide. This chapter discusses frequency bands, modulation such as quadrature PSK and adaptive differential pulse code modulation, geosynchronous and geostationary orbits, and different types of antennas such as primary focus, Cassegrain, and Gregorian. Noise, equivalent temperature, and gain over temperature are discussed as well as the method used to evaluate different systems and the link budget. Multiple channels and multiple-access techniques are discussed for increased capacity. Propagation delay; the cost of use depending on the types of transmission, which includes permanently assigned multiple-access PAMA, demand assigned multiple-access DAMA, or occasional; and the different types of satellites used for communications

are addressed, including low earth orbit satellites, geosynchronous earth orbit satellites, and medium earth orbit satellites.

Chapter 13 discusses the global positioning system (GPS), which uses a direct sequence spread-spectrum binary PSK data link. This chapter includes coarse acquisition code, precision code, data signal structure, receiver characteristics, errors due to the atmosphere, multipath, Doppler, and selective availability, which has been turned off. It also discusses the pros and cons of using narrow correlation, carrier smoothing of the code (integrated Doppler), differential GPS, and relative GPS. Kinematic carrier phase tracking, the double difference, and wide lane versus narrow lane techniques are also discussed. In addition, other satellite positioning systems are discussed, including the Global Navigational Satellite System from the Soviet Union commonly known as GLONASS, and the Galileo In-Orbit Validation Element in Europe.

Chapter 14 covers fundamentals of Radar and Radar communications. It discusses the Radar applications, types of Radar used with most of the focus on pulse Radar modulation and operation, the Radar cross section, and developing the Radar path budget to arrive at the basic Radar equations. This also includes the determination and calculations for range equation and compares it to a sound wave example. This chapter discusses range ambiguity, range resolution, bearing and angle resolution and Radar accuracy, linear and angular velocity, and the display results using a PPI and A-scope. This covers types of Radar antennas, types of Radars including search, acquisition and track Radars, missile and missile guidance Radars, airborne and frequency diversity Radars, chirped and digital pulse compression Radars, frequency-modulated continuous wave FMCW Radars, Doppler Radars including Doppler measurement, passive and active Radars, synthetic aperture SAR Radars, next-generation Weather Radars, and terminal Doppler Weather Radar TDWR. Radar design including transmitter, receiver, and pulse shaping is discussed. Other topics covered are minimum discernable signal MDS, clutter, frequency bands used, moving target indication MTI, blind speeds, sampling process, and multiple-pulse MTI Radar. In addition, this chapter covers Radar communications which involves using the Radar for communications. This can be accomplished by using direct burst pulse-coded modulation or pulse position modulation or integrating both methods. Probability of detect and false alarm are also discussed for a radar pulse and system trade-offs. Including is a discussion for Radar communication applications and a discussion of the advantages of PPM including minimizing Doppler effects. Chapter 15 discusses direction finding and interferometer analysis using direction cosines and coordinate conversion techniques to provide the correct solution. This chapter provides information on the limitations of the standard interferometer equation and details the necessary steps to design a three-dimensional interferometer solution for the yaw, pitch, and roll of antennas.

Each of the chapters contains problems, with answers provided in the back of the book along with appendices for further information.

If you have any suggestions, corrections, or comments, please send them to Scott Bullock at scottrbullock@gmail.com.

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About the author

Scott R. Bullock received his BSEE degree from Brigham Young University in 1979 and his MSEE degree from the University of Utah in 1988. Mr. Bullock worked in research and development for most of his career developing a radar simulator, a spread-spectrum Microscan receiver, and a new spread-spectrum receiver, for which he applied for a patent and was awarded company funds as a new idea project to develop the concept. Mr. Bullock also developed a spread-spectrum environment simulator for a spread-spectrum wideband countermeasures receiver using binary phase-key shifting, quadrature PSK, offset quadrature PSK, minimum-shift keying, frequency hopper, hybrids, amplitude modulation, frequency modulation, voice generator, jammers, and noise. He also designed a high-frequency adaptive filter used to reduce narrowband jammers in a wideband signal; a broadband, highly accurate frequency hop detector; an instantaneous Fourier transform receiver; a chopper modulated receiver; a Ku-band radio design for burst spread-spectrum communications through a Troposcatter channel; a GSO to reduce jammers; an advanced tactical data link; radio frequency analysis of an optical receiver study; a portable wideband communications detector; and an acoustic-optic spectrum analyzer photodiode array controller, and several ground-to-ground, ground-to-air, air-to-air, ground-to-satellite, and satellite-to-satellite communication links and data collection.

Mr. Bullock developed the first-handheld PCS spread-spectrum telephone with Omnipoint in the 902–928 MHz ISM band and later in the 2.4-GHz ISM band. He also received a patent for his work on reducing spectral lines to meet the Federal Communications Commission power spectral density requirements. He developed the wireless data link to replace the wires in the TOW missile using CP-PSK modulation.

He was responsible for various types of spread-spectrum data links for the SCAT-1 program related to aircraft GPS landing systems. He was an active participant in the RTCA meetings held in Washington D.C. for the evaluation and selection of the D8PSK data link to be used as the standard in all SCAT-1 systems. He also worked on the concepts of the Wide Area Augmentation System, low probability of intercept data link, rapid assembly of Army forces using relative GPS and a spread-spectrum data link, DS/FH air traffic control asynchronous system, JTRS, and Link-16.

Mr. Bullock developed several commercial products such as wireless jacks for telephones, PBXs, modems, wireless speakers, and other various wireless data link products.

He has designed directional volume search, tracking algorithms, and cognitive systems and networks for both the commercial and military communities. In addition, he designed, assembled, and successfully field tested a network of common data links from multiple vendors using a multibeam active electronic steerable array AESA.

He has worked in the area of satellites developing ideal waveforms for satellite networks and communications, radar communications using PCM and PPM, differential systems to mitigate Doppler and oscillator drifts, and developed a CP-PSK ideal waveform for two-way missile control and communications, two-way SATCOM, and several other applications for spectral efficiency, reduction in sidelobe regeneration, and immunity to saturation effects.

Mr. Bullock has held many high-level positions, such as vice president of engineering for Phonex Broadband, vice president of engineering for L-3 Satellite Network Division, senior director of engineering for MKS/ENI, engineering fellow for Raytheon, engineering manager/consulting engineer for Northrop Grumman, and senior system/architect engineer for General Dynamics. He specializes in wireless data link design, cognitive systems, and system analysis and directs the design and development of wireless products for both commercial and military customers.

Mr. Bullock holds 19 patents and 44 trade secrets in the areas of spread-spectrum wireless data links, adaptive filters, frequency hop detectors, wireless telephone, and data products. He has published several articles dealing with spread-spectrum modulation types, multipath, AGCs, PLLs, and adaptive filters. He is the author of this book and another book titled *Broadband Communications and Home Networking*. He is a licensed professional engineer and a member of IEEE and Eta Kappa Nu, and he holds an Extra Class Amateur Radio License, KK7LC.

He has performed data link communications work and taught multiple seminars and classes for Texas Instruments, L-3comms, BAE, Omnipoint, E-Systems, Raytheon, Phonex Broadband, Northrop Grumman, General Dynamics, SPAWAR, NASA, NAVAIR, CIA, SAIC, MKS/ENI, Thales, and the IEEE Smart Tech Metro Area Workshops in both Baltimore and Atlanta. He is currently an instructor for Besser Associates, ATI courses, and K2B International. He taught the advanced communication course at ITT, an engineering course at PIMA Community College, and was a guest lecturer on multiple-access systems at PolyTechnic University, Long Island, New York.

Chapter 1
Transceiver design

A transceiver is a system that contains both a transmitter and a receiver. The transmitter from one transceiver sends a signal through space to the receiver of a second transceiver. After receiving the signal, the transmitter from the second transceiver sends a signal back to the receiver of the first transceiver, completing a two-way communications data link system, as shown in Figure 1.1.

There are many factors to consider when designing a two-way communications link. The first one is to determine the operating frequency. Several considerations need to be evaluated to select the frequency that is going to be used.

1.1 Frequency of operation

In a transceiver design, the first step is to determine the radio frequency (RF) of operation. The frequency of operation depends on the following factors:

- RF availability: This is the frequency band that is available for use by a particular system and is dependent on the communications authority for each country. For example, in the United States, it is specified by the Federal Communications Commission (FCC), and in the United Kingdom, it is specified by the British Approvals Board for Telecommunications. These two groups have ultimate control over frequency-band allocation. Other organizations that help to establish standards are the International Telecommunications Union Standardization Sector, the European Conference of Postal and Telecommunications Administrations, and the European Telecommunications Standards Institute. In addition, each country may have their own communication authorities, and the frequency selection will need to get the approvals for that country.
- Cost: As the frequency increases, the components in the receiver tend to be more expensive. An exception to the rule is when there is a widely used frequency band, such as the cellular radio band, where supply and demand drives down the cost of parts and where integrated circuits are designed for specific applications. These are known as application-specific integrated circuits.
- Range and antenna size: Increasing the frequency increases the loss between the transmitter and the receiver which reduces the range of data link. The main contributor of this loss is called Free-space Loss and is calculated using the

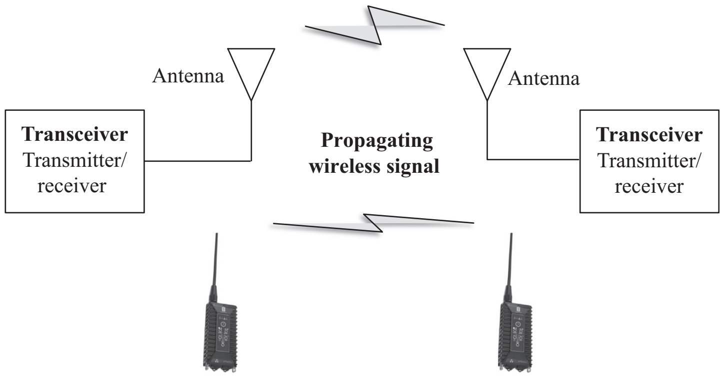


Figure 1.1 Two-way wireless transceiver block diagram

frequency or wavelength of the transmission. This results in a decrease in range for line-of-sight applications or an increase in the output power requirement, which would also affect cost.

- Reflections: Another way to increase range using lower operational frequencies is called skip or reflection off the atmosphere, mainly the ionosphere and sometimes the troposphere. This communications is not limited to line-of-site and can travel much further distances by using multiple reflections.

For specific frequencies, this can increase the range tremendously. Amateur radio operators use frequencies that can reflect off the atmosphere and travel half way around the world with less than 100 W of power.

The size of the antenna decreases as the frequency increases. Lower frequencies improve the range; however, the size of the antenna might be too big for practical considerations and could also be a factor in the cost of the design. For high frequencies, the antenna size is small and less expensive; however, the range is decreased. These trade-offs need to be made to ensure the best selection of the frequency of operation.

- Customer specified: Often times, the frequency of operation is specified by the customer. If the application is for commercial applications, the frequency selection must follow the rules currently in place for that specific application to obtain the approval of the FCC and other agencies.
- Band congestion: Ideally, the frequency band selected is in an unused band, especially with no high-power users in the band. Generally, the less-used bands are very high frequencies that increase the cost. Many techniques available today allow more users to operate successfully in particular bands, and some of these techniques will be discussed further in the book. Whatever band is selected, it requires approval by the FCC or other agencies.

The frequency of operation is then selected by taking into consideration the aforementioned criteria.

A listing of the basic frequency bands is shown in Table 1.1 with some applications specified. More detailed frequency allocations can be obtained from the FCC website or in the literature.

The frequency bands are all allocated to different users, which makes it virtually impossible to obtain a band that is not already allocated. However, some band reorganizing and renaming have occurred; for example, some of the old existing analog television bands have been reallocated to use for multiple digital wireless applications.

In addition, there are basically two accepted frequency-band designations using letters of the alphabet, and these are listed in Table 1.2. This can cause some confusion when designating the band of operation. For example, both frequency-band designations include the K band with different frequency ranges.

Even though a new designation of the frequency bands has been introduced, the old legacy designation is still commonly used today.

Table 1.1 Frequency bands

- **ELF** Extremely low frequency 0–3 kHz
- **VLF** Very low frequency 3–30 kHz
- **LF** Low frequency 30–300 kHz
- **MF** Medium frequency 300–3,000 kHz—AM radio broadcast
- **HF** High frequency 3–30 MHz—Shortwave broadcast radio
- **VHF** Very high frequency 30–300 MHz—TV, FM radio broadcast, mobile/fixed radio
- **UHF** Ultra-high frequency 300–3,000 MHz—TV
- **L band** 500–1,500 MHz—PCS/cell phones
- **ISM bands** 902–928 MHz, 2.4–2.483 GHz, 5.725–5.875 GHz PCS and RFID
- **S band** 2–4 GHz—Cell phones
- **C band** 3,600–7,025 MHz—Satellite communications, radios, radar
- **X band** 7.25–8.4 GHz—Mostly military communications
- **Ku band** 10.7–14.5 GHz—Satellite communications, radios, radar
- **Ka band** 17.3–31.0 GHz—Satellite communications, radios, radar
- **SHF** Super high frequencies—(Microwave) 3–30.0 GHz
- **EHF** Extremely high frequencies—(Millimeter wave signals) 30.0–300 GHz—Satellite
- **Infrared radiation** 300–430 THz (Terahertz)—infrared applications
- **Visible light** 430–750 THz
- **Ultraviolet radiation** 1.62–30 PHz (Petahertz)
- **X-rays** 30 PHz–30 EHz (Exahertz)
- **Gamma rays** 30–3,000 EHz

Table 1.2 Frequency-band designations

Legacy Designations		New Designations	
Band	**Frequency range (GHz)**	**Band**	**Frequency range (GHz)**
I band	to 0.2	A band	to 0.25
G band	0.2–0.25	B band	0.25–0.5
P band	0.25–0.5	C band	0.5–1.0
L band	0.5–1.5	D band	1–2
S band	2–4	E band	2–3
C band	4–8	F band	3–4
X band	8–12	G band	4–6
K_u band	12–18	H band	6–8
K band	18–26	I band	8–10
K_a band	26–40	J band	10–20
V band	40–75	K band	20–40
W band	75–111	L band	40–60
		M band	60–100

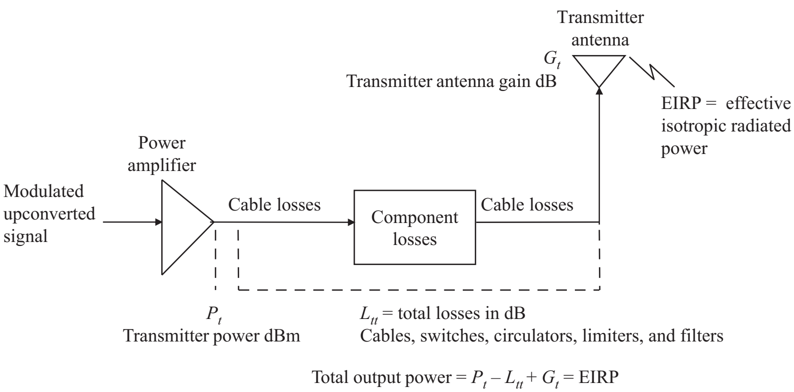


Figure 1.2 RF section of the transmitter

1.2 Transmitter

The transmitter is the part of the transceiver that creates, modulates, and transmits the signal through space to the receiver. The transmitter is responsible for providing the power required to transmit the signal through the link to the receiver. This includes the power amplifier (PA), the transmitter antenna, and the gains and losses associated with the process, such as cable and component losses, to provide the effective isotropic radiated power (EIRP) out of the antenna (Figure 1.2).

1.2.1 Transmitted effective isotropic radiated power

An isotropic radiator is a theoretical radiator that assumes a point source radiating in all directions. EIRP is the amount of power from a single point radiator that is

required to equal the amount of power that is transmitted by the PA, losses, and directivity of the antenna (antenna gain) in the direction of the receiver.

The EIRP provides a way to compare different transmitters. To analyze the output of an antenna, EIRP is used (Figure 1.2):

$$\text{EIRP} = P_t - L_{tt} + G_t - L_{ta}$$

where P_t is the transmitter power in dBm, L_{tt} is the total losses from the PA to the antenna in dB; coaxial or waveguide line losses, switchers, circulators, antenna connections, G_t is the transmitter antenna gain in dB referenced to a isotropic antenna, and L_{ta} is the total transmitter antenna losses in dB.

Effective radiated power (ERP) is another term used to describe the output power of an antenna. However, instead of comparing the effective power to an isotropic radiator, the power output of the antenna is compared to a dipole antenna. The relationship between EIRP and ERP is:

$$\text{EIRP} = \text{ERP} + G_{\text{dipole}},$$

where G_{dipole} is the gain of a dipole antenna, which is equal to approximately 2.14 dB (Figure 1.2). For example,

$$\begin{aligned} \text{EIRP} &= 10\,\text{dBm} \\ \text{ERP} &= \text{EIRP} - G_{\text{dipole}} = 10\,\text{dBm} - 2.14\,\text{dB} = 7.86\,\text{dBm} \end{aligned}$$

1.2.2 Power from the transmitter

The power from the transmitter (P_t) is the amount of power output of the final stage of the PA. For ease in the analysis of power levels, the power is specified in dBm or converted to dBm from milliwatts (mW). The power in mW is converted to power in dBm by:

$$P_{\text{dbm}} = 10 \log P_{\text{mW}}$$

Therefore, 1 mW is equal to 0 dBm. The unit dBm is used extensively and a good understanding of this term and other dB terms is important. dBm is the most common expression of power in the communications industry. The term dBm is actually a power level related to 1 mW and is not a loss or gain as is the term dB.

A decibel (dB) is a unit for expressing the ratio of two amounts of electric or acoustic signal power. The decibel is used to enable the engineer to calculate the resultant power level by simply adding or subtracting gains and losses instead of multiplying and dividing. Also it makes it simple to work with very small power levels.

Gains and losses are expressed in dB. A dB is defined as a power ratio:

$$\text{dB} = 10 \log(P_o/P_i)$$

where P_i is the input power (in mW) and P_o is the output power (in mW).

For example,

Given:

Amplifier power input = 0.15 mW = 10 log(0.15) = −8.2 dBm
Amplifier power gain $P_o/P_i = 13 = 10 \log(13) = 11.1$ dB

Calculate the power output:

Power output = 0.15 mW × 13 = 1.95 mW using power and multiplication
Power output (in dBm) = −8.2 dBm + 11.1 dB = 2.9 dBm using dBm and dB and addition
Note: 2.9 dBm = 10 log(1.95)

Another example of using dBm and dB is as follows:

$P_i = 1$ mW or 0 dBm
Attenuation = 40 dB
Gain = 20 dB
$P_o = 0$ dBm − (40 dB attenuation) + (20 dB gain) = −20 dBm

In many applications, dB and dBm are misused, which can cause errors in the results. The unit dB is used for a change in power level, which is generally a gain or a loss. The unit dBm is used for absolute power; for example, 10 log(1 mW) = 0 dBm. The unit dBW is also used for absolute power in Watts 10 log(1 W) = 0 dBW. The terms dBm and dBW are never used for expressing a change in signal level. The following examples demonstrate this confusion.

Example 1: Suppose that there is a need to keep the output power level of a receiver at 0 dBm ±3 dB. If ±3 dB is mistakenly substituted by ±3 dBm in this case, the following analysis would be made:

Given:

0 dBm = 1 mW (power level in milliwatts)
3 dBm = 2 mW (power level in milliwatts)
3 dB = 2 (or two times the power or the gain)
−3 dB = 1/2 (half the power or the loss)

The erroneous analysis using ±3 dBm would be as follows:

0 dBm ± 3 dBm
0 dBm + 3 dBm = 1 mW + 2 mW = 3 mW; 4.8 dBm; the wrong answer
0 dBm − 3 dBm = 1 mW − 2 mW = −1 mW; cannot have negative power

The correct analysis using ±3 dB would be as follows:

0 dBm ± 3 dB
0 dBm + 3 dB = 1 mW × 2 = 2 mW = 3 dBm; correct answer
0 dBm − 3 dB = 1 mW × 1/2 = 0.5 mW = −3 dBm; correct answer

Using the correct form:

0 dBm + 3 dB = 3 dBm
0 dBm − 3 dB = −3 dBm

Therefore, they are not interchangeable; the incorrect answers are the result of using the wrong term. Remember, dBm represents an actual power level, and dB represents a change in power level, gain, or loss.

Example 2: Figure 1.3 shows a graph of a voltage-controlled amplifier, with the control voltage on the *x* axis and power output (in dBm) on the *y* axis. The slope is equal to:

$$\begin{aligned}\text{Slope} &= \text{Rise/Run} = [(-40\ \text{dBm}) - (-60\ \text{dBm})]/(5\ \text{V} - 3\ \text{V})\\ &= 20/2 = 10\ \text{dB/V}, \text{not dBm/V}\end{aligned}$$

This example is showing a change in power level, which is specified in dB.

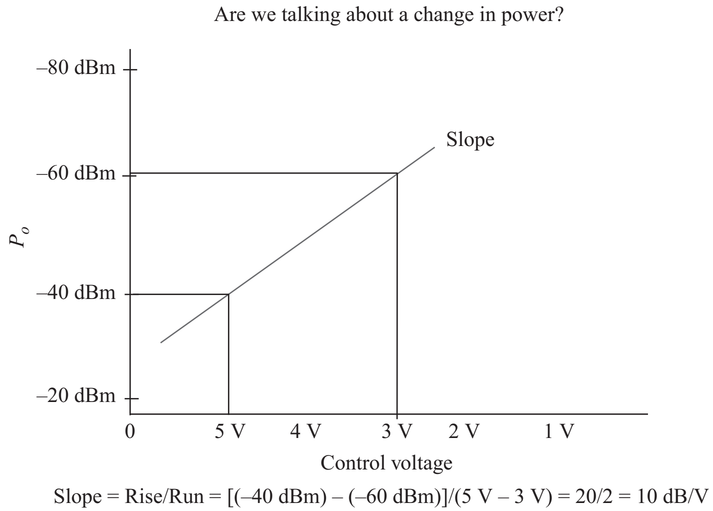


Figure 1.3 Voltage-controlled amplifier curve

1.2.2.1 Voltage and power using dB

The term dB can be described both in power and voltage. For voltage, the same equation for converting losses and gains into dB is used, substituting V^2/R for P:

$$P_i = V_i^2/R_i$$
$$P_o = V_o^2/R_o$$
$$\text{dB} = 10\log\frac{(P_o)}{(P_i)} = 10\log\frac{\left(V_0^2/R_o\right)}{\left(V_i^2/R_i\right)}$$

where P_o is the power out, P_i is the power in, V_o is the voltage out, V_i is the voltage in, R_o is the output impedance, and R_i is the input impedance.

If $R_o = R_i$, they cancel, and the resultant gain/loss equation is:

$$\text{dB} = 10\log\frac{V_o^2}{V_i^2} = 10\log\left(\frac{V_o}{V_i}\right)^2 = 20\log\frac{V_o}{V_i}$$

Assuming $R_o = R_i$, dB is the same whether voltage or power is used. Therefore, if the system has 6 dB of gain, it has 6 dB of voltage gain and 6 dB of power gain. The only difference is that the ratio of voltage is increased by two and the ratio of power is increased by four.

For example, if an amplifier has 6 dB of gain, it has four times more power and two times more voltage at the output of the amplifier referenced to the input. Therefore, if the input to the amplifier is 5 V and the power at the input is 0.5 W, the output voltage would be 10 V (two times) and the output power would be 2 W (four times) given that $R_i = R_o$. So the gain would be:

$$\text{Voltage gain} = 20\log(V_o/V_i) = 20\log(10\text{ V}/5\text{V}) = 20\log(2) = 6\ \text{dB}$$
$$\text{Power gain} = 10\log(P_o/P_i) = 10\log(2/0.5) = 10\log(4) = 6\text{ dB}$$

If $R_i \neq R_o$, then the dB gain for voltage is different from the dB gain for power. If only voltage gain is desired, dBv is used indicating dB volts. This voltage gain can be expressed in millivolts (mV), or dBmv, indicating dB millivolts (mV). For example, going from 10 to 20 mV or doubling the voltage which would be 6 dBmv of gain. However, the caution is that simply changing the ratio R_i/R_o can change the voltage gain without an increase in power gain. For example,

Given:

$$V_i = 1\text{ V}, R_i = 1\text{ ohm}, V_o = 2\text{ V}, R_o = 4\text{ ohm}$$
$$P_i = V_i^2/R_i = 1^2/1 = 1\text{ W}$$
$$P_o = V_o^2/R_o = 2^2/4 = 1\text{ W}$$

Then,

Power gain in dB: $10\log(P_o/P_i) = 10\log(1/1) = 10\log(1) = 0$ dB power gain
Voltage gain in dB: $20\log(V_o/V_i)$ $20\log(2/1) = 20\log(1/2) = 6$ dBv voltage gain

Since $R_i \neq R_o$, the voltage and power gain are different.

There is no power gain in the previous example, but by changing the resistance, a voltage gain of 6 dBv was realized. Power is equal to the voltage times the current ($P = VI$). If the voltage has a gain of 2, the current needs a gain of 1/2 in order to keep the power unchanged.

The term dB is a change in signal level, signal amplification, or signal attenuation. It is the ratio of power output to power input. It can also be the difference from a given power level, such as "so many dB below a reference power."

The following are some definitions for several log terms that are referred to and used in the industry today:

- dB is the ratio of signal levels, which indicates the gain or loss in signal level.
- dBm is a power level in milliwatts: 10 log P, where $P =$ power (in mW).
- dBW is a power level in watts: 10 log P, where $P =$ power (in W).
- dBc is the difference in power between a signal that is used as the carrier frequency and another signal, generally an unwanted signal, such as a harmonic of the carrier frequency.
  - Carrier power = 0 dBm, 5th harmonic is −60 dBm
  - 5th harmonic is 60 dBc down from carrier or −60 dBc
- dBi is the gain of an antenna with respect to the gain of an ideal isotropic radiator.
  - A dipole antenna is ~2.14 dBi

Several other dB terms, similar to the last two, refer to how many dB away a particular signal is from a reference level. For example, if a signal's power is 20 times smaller than a reference signal's power *r*, then the signal is −13 dBr, which means that the signal is 13 dB smaller than the reference signal.

There is one more point to consider when applying the term dB. When referring to attenuation or losses, the output power is less than the input power, so the attenuation in dB is subtracted instead of added. For example, if we have a power of +5 dBm and the attenuation in dB is equal to 20 dB, then the output of the signal level is equal to:

$$+5\,\text{dBm} - 20\,\text{dB losses} = -15\,\text{dBm}$$

If the attenuator is specified with a minus sign, then it is added in the equation giving the same results.

Last, if two power levels are specified, the gain or loss can be calculated by adding or subtracting the power levels as:

- 0 dBm + 3 dBm = 3 dB gain
- 0 dBm − 3 dBm = −3 dB loss

1.2.3 Transmitter component losses

Most transceiver systems contain RF components such as a circulator or a transmit/receive (T/R) switch that enables the transceiver to use the same antenna for both transmitting and receiving. Also, if the antenna arrays use multiple antennas, some of their components will interconnect the individual antenna elements. Since these elements have a loss associated with them, they need to be taken into account in the overall output power of the transmitter. These losses directly reduce the signal level or the power output of the transmitter. The component losses are labeled and are included in the analysis:

$$L_{t\text{comp}} = \text{switchers; circulators; antenna connections}$$

Whichever method is used, the losses directly affect the power output on a one-for-one basis. A 1-dB loss equals a 1-dB loss in transmitted power. Therefore, the

losses after the final output PA of the transmitter and the first amplifier [or low-noise amplifier (LNA)] of the receiver should be kept to a minimum. Each dB of loss in this path will either reduce the minimum detectable signal by a dB or the transmitter gain will have to transmit a dB more power.

1.2.4 Transmitter line losses from the power amplifier to the antenna

Since most transmitters are located at a distance from the antenna, the cable or waveguide connecting the transmitter to the antenna contains losses that need to be incorporated in the total power output:

$$L_{tll} = \text{coaxial or waveguide line losses (in dB)}$$

These transmitter-line losses are included in the total power output analysis; a 1-dB loss equals a 1-dB loss in power output. Using larger diameter cables or higher quality cables can reduce the loss, which is a trade-off with cost. For example, Heliax cables are used for very low-loss applications. However, they are generally more expensive and larger in diameter than standard cables. The total losses between the PA and the antenna are therefore equal to:

$$L_{tt} = L_{tll} + L_{t\text{comp}}$$

Another way to reduce the loss between the transmitter and the antenna is to locate the transmitter PA as close to the antenna as possible. This will reduce the length of the cable, which reduces the overall loss in the transmitter.

1.2.5 Transmitter antenna gain

Most antennas experience gain because they tend to focus energy in specified directions compared with an ideal isotropic antenna, which radiates in all directions. A simple vertical dipole antenna experiences approximately 2.14 dBi of gain compared with an isotropic radiator because it transmits most of the signal around the antenna, with very little of the signal transmitted directly up to the sky and directly down to the ground (Figure 1.4).

Antennas do not amplify the signal power but focus the existing signal in a given direction. This is similar to a magnifying glass, which can be used to focus the sun rays in a specific direction, increasing the signal level at a single point (Figure 1.5).

A parabolic dish radiator is commonly used at high frequencies to achieve gain by focusing the signal in the direction the antenna is pointing (Figure 1.5). The gain for a parabolic antenna is

$$G_t = 10 \log\left[n(\pi D/\lambda)^2\right]$$

where G_t is the gain of the antenna (in dBi), n is the efficiency factor <1, D is the diameter of the parabolic dish, and λ is the wavelength.

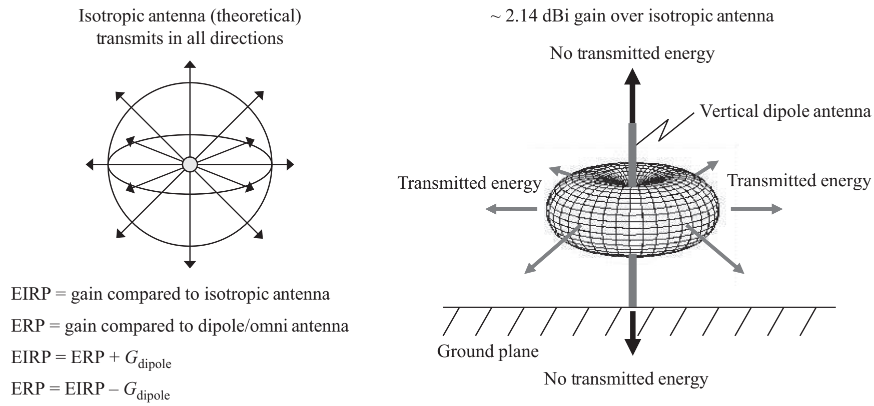


Figure 1.4 A simple vertical dipole antenna gain compared to an isotropic radiator

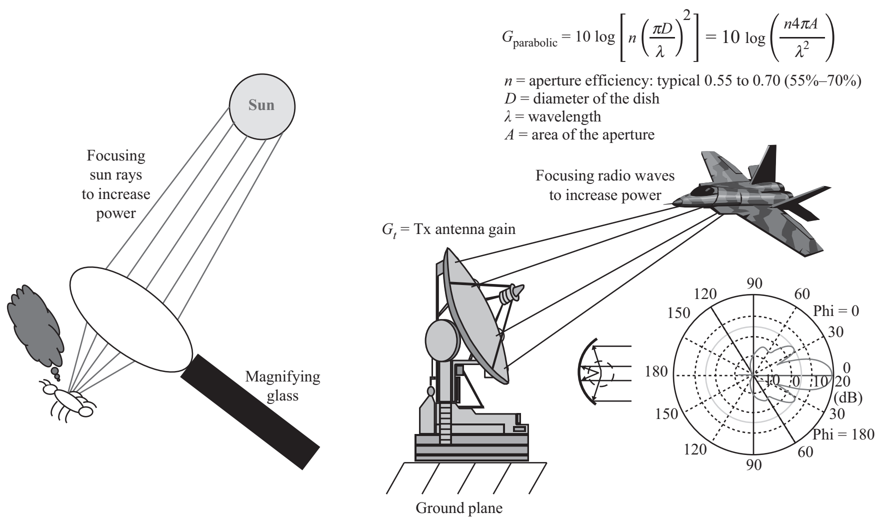


Figure 1.5 Focusing increases power which provides gain

The efficiency factor is the actual gain of the antenna compared with the theoretical gain. This can happen when a parabolic antenna is not quite parabolic, when the surface of the antenna combined with the feed is not uniform, and when other anomalies occur in the actual implementation of the antenna system. Typically, this ranges from 0.5 to 0.8, depending on the design and the frequency of operation.

Notice that the antenna gain increases both with increasing diameter and higher frequency (shorter wavelength). The gain of the antenna is a direct gain

where a 1-dB gain equals a 1-dB improvement in the transmitter power output. Therefore, a larger gain will increase the range of the link. In addition, the more gain the antenna can produce, the less power the PA has to deliver for the same range. This is another trade-off that needs to be considered to ensure the best design and the lowest cost for a given application.

1.2.6 Transmitter antenna losses

Several losses are associated with the antenna. Some of the possible losses, which may or may not be present in each antenna, are as follows:

- L_{tr}, radome losses on the transmitter antenna. The radome is the covering over the antenna that protects the antenna from the outside elements. Most antennas do not require a radome.
- $L_{t\mathrm{pol}}$, polarization mismatch losses. Many antennas are polarized (i.e., horizontal, vertical, or circular). This defines the spatial position or orientation of the electric and magnetic fields. A mismatch loss is due to the polarization of the transmitter antenna being spatially off with respect to the receiver antenna. The amount of loss is equal to the angle difference between them. For example, if both the receiver and transmitter antennas are vertically polarized, they would be at 90° from the earth. If one is positioned at 80° and the other is positioned at 100°, the difference is 20°. Therefore, the loss due to polarization would be

$$20\log(\cos\ \theta) = 20\log(\cos 20^\circ) = 0.54\ \mathrm{dB}$$

- $L_{t\mathrm{foc}}$, focusing loss or refractive loss. This is caused by imperfections in the shape of the antenna so that the energy is focused toward the feed. This is often a factor when the antenna receives signals at low elevation angles.
- $L_{t\mathrm{point}}$, mispointed loss. This is caused by transmitting and receiving directional antennas that are not exactly lined up and pointed toward each other. Thus, the gains of the antennas do not add up without a loss of signal power.
- $L_{t\mathrm{con}}$, conscan crossover loss. This loss is present only if the antenna is scanned in a circular search pattern, such as a conscan (conical scan) radar searching for a target. Conscan means that the antenna system is either electrically or mechanically scanned in a conical fashion or in a cone-shaped pattern. This is used in radar and other systems that desire a broader band of spatial coverage but must maintain a narrow beam width. This is also used for generating the pointing error for a tracking antenna.

The total transmitter antenna losses are:

$$L_{ta} = L_{tr} + L_{t\mathrm{pol}} + L_{t\mathrm{foc}} + L_{t\mathrm{point}} + L_{t\mathrm{con}}$$

These losses are also a direct attenuation: a 1-dB loss equals a 1-dB loss in the transmitter power output.

1.3 The channel

The channel is the path of the RF signal that is transmitted from the transmitter antenna to the receiver antenna. This is the signal in space that is attenuated by the channel medium. The main contributor to channel loss is free-space attenuation. The other losses, such as the propagation losses and multipath losses are fairly small compared with free-space attenuation. The losses are depicted in Figure 1.6 and are described in detail in the following sections.

1.3.1 Free-space attenuation

As a wave propagates through space, there is a loss associated with it. This loss is due to dispersion, the ‘‘spreading out’’ of the beam of radio energy as it propagates through space. This loss is consistent and relative to wavelength, which means that it increases with frequency as the wavelength becomes shorter. This is called free-space loss or path loss and is related to both the frequency and the slant range, which is the distance between the transmitter and receiver antennas. Free-space loss is given by:

$$A_{fs} = 20\ \log[4\pi R/\lambda] = 20\ \log[4\pi Rf/c]$$

where A_{fs} is the free-space loss, R is the slant range (same units as λ), λ is the wavelength $= c/f$, f is the frequency of operation, c is the speed of light, 300×10^6 m/s, and R is in meters.

Therefore, the free-space loss increases as both the range and frequency increase. This loss is also a direct attenuation: a 1-dB loss equals a 1-dB loss in the analysis.

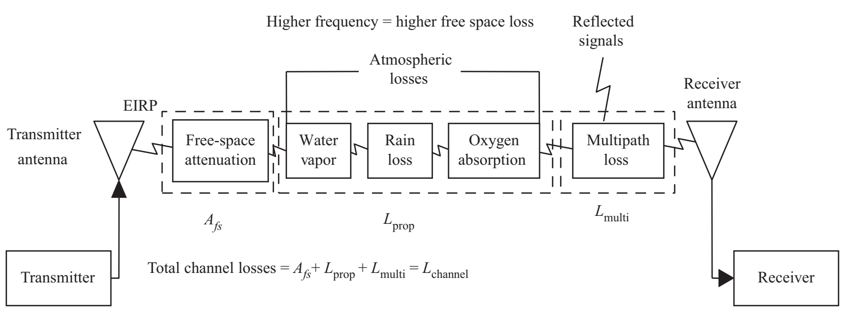


Figure 1.6 Channel losses for link budget

There are many forms of free-space loss depending on the functions in the equation;

$A_{fs} = (\lambda/(4\pi))^2$ will be less than 1 and multiplied
$A_{fs} = ((4\pi R)/\lambda)^2$ will be greater than 1 and divided
$A_{fs\mathrm{dB}} = 10\log(\lambda/(4\pi R))^2 = 20\log 1/(4\pi R) =$ will be a negative number and added
$A_{fs\mathrm{dB}} = 10\log((4\pi R)/\lambda)^2 = 20\log(4\pi R)/1 =$ will be a positive number and subtracted

It is important to determine if it is multiplied or divided, added or subtracted, to avoid mistakes and possible confusion.

Some examples that use the different forms of free-space loss are shown below:

Given: $P_t = 100$ W $= 50$ dBm, $\lambda = 0.125$, $R = 1{,}000$ m
$A_{fs} = (\lambda/(4\pi R))^2 = 98.9 \times 10^{-12}$ Multiply: $P_r = 100\ \mathrm{W} \times 98.9 \times 10^{-12} = 9.89 \times 10^{-9}\ \mathrm{W} = 9.89 \times 10^{-6}$ mW
$A_{fs} = (4\pi R)/\lambda)^2 = 1.01065 \times 10^{10}$ Divide: $P_r = 100\ \mathrm{W}/(1.01065 \times 10^{10}) = 9.89 \times 10^{-9}\ \mathrm{W} = 9.89 \times 10^{-6}$ mW
$A_{fs} = 20\log(\lambda/(4\pi R) = -100$ dB Add: $P_r = 50\ \mathrm{dBm} + (-100\ \mathrm{dB}) = -50$ dBm
$A_{fs} = 20\log(4\pi R)/\lambda = 100$ dB Subtract: $P_r = 50\ \mathrm{dBm} - 100\ \mathrm{dB}) = -50$ dBm
Note: $10\log(9.89 \times 10^{-6}\ \mathrm{mW}) = -50$ dBm

Even though free-space attenuation is treated as a loss, free space does not attenuate the signal depending on frequency. It is caused by electromagnetic energy being spread with respect to distance, and the size of the Rx antenna depends on wavelength/frequency. The larger the antenna, the more power it receives since it has a larger service area to collect the power.

1.3.2 Propagation losses

There are three main losses depending on the conditions in the atmosphere; clouds, rain, and humidity:

- Cloud loss, loss due to water vapor in clouds.
- Rain loss, loss due to rain.
- Atmospheric absorption (oxygen loss), due to the atmosphere.

These values vary from day-to-day and from region to region. Each loss depends on the location, and a nominal loss, generally not the worst case, is used for calculating the power output required from the transmitter.

To get a feel for the types of attenuation that can be expected, some typical numbers are as follows:

- Rain loss $= -1$ dB/mi at 14 GHz with 0.5″/h rainfall
- Rain loss $= -2$ dB/mi at 14 GHz with 1″/h rainfall
- Atmospheric losses $= -0.1$ dB/nautical mile for a frequency range of approximately 10–20 GHz

- Atmospheric losses = −0.01 dB/nautical mile for a frequency range of approximately 1–10 GHz

A typical example gives a propagation loss of approximately −9.6 dB for the following conditions:

- Frequency = 18 GHz
- Range = 10 nm
- Rainfall rate = 12 mm/h
- Rain temperature = 18°C
- Cloud density = 0.3 g/m^3
- Cloud height = 3–6 km

To determine the actual loss for a particular system, several excellent sources can be consulted. These propagation losses are also a direct attenuation for calculating the transmitter power output: a 1-dB loss equals a 1-dB loss in the power output.

1.3.3 Multipath losses

Whenever a signal is sent out in space, the signal can either travel on a direct path from the transmitter antenna to the receiver antenna or take multiple indirect paths caused by reflections off objects, which is known as multipath. The most direct path the signal can take has the least amount of attenuation. The other paths (or multipath) are attenuated but can interfere with the direct path at the receiver. The effect of multipath can cause large variations in the received signal strength. Multipath loss is incorporated into the link budget and is considered a loss in signal level. This loss is a direct attenuation: a 1-dB loss equals a 1-dB loss in the link analysis.

1.4 Receiver

The receiver accepts the signal that was sent by the transmitter through the channel via the receiver antenna (Figure 1.7). The losses from the antenna to the LNA, which is the first amplifier in the receiver, should be kept as small as possible. This includes cable losses and component losses that are between the antenna and the LNA (Figure 1.7). The distance between the receiver antenna and the LNA should also be as small as possible. Some systems actually include the LNA with the antenna system, separate from the rest of the receiver, to minimize this loss. The main job of the receiver is to receive the transmitted signal and detect the data that it carries in the most effective way without further degrading the signal.

1.4.1 Receiver antenna losses

Antenna losses for the receiver are very similar to those for the transmitter, some of which are listed as follows:

- L_{rr}, radome losses on the receiver antenna. The radome is the covering over the antenna that protects the antenna from the outside elements.

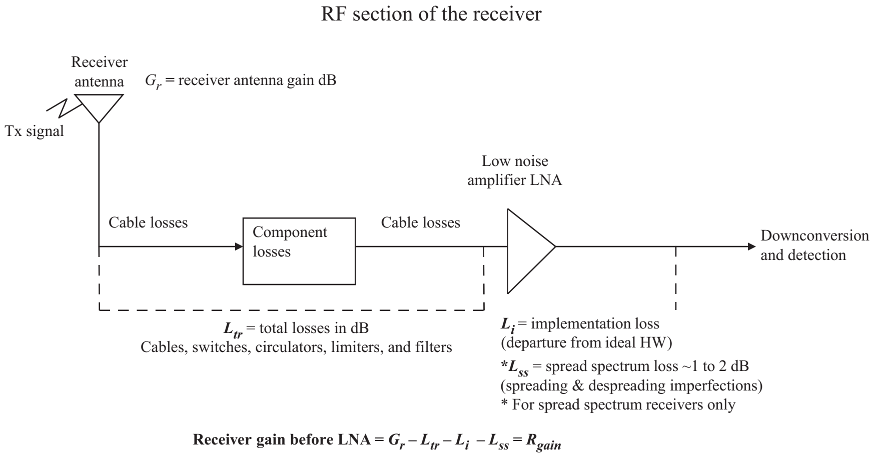


Figure 1.7 RF section of the receiver

- $L_{r\text{pol}}$, polarization loss. Many antennas are polarized (i.e., horizontal, vertical, or circular). This defines the spatial position or orientation of the electric and magnetic fields.
- $L_{r\text{foc}}$, focusing loss or refractive loss. This is a loss caused by imperfections in the shape of the antenna, so that the energy is focused toward the feed. This is often a factor when the antenna receives signals at low elevation angles.
- $L_{r\text{point}}$, mispointing loss. This is caused by transmitting and receiving directional antennas that are not exactly lined up and pointed toward each other. Therefore, the gains of the antennas do not add up without a loss of signal power. Note that this loss may be combined into one number so that it is not included in both the receiver analysis and the transmitter analysis.
- $L_{r\text{con}}$, conscan crossover loss. This loss is present if the antenna is scanned in a circular search pattern.

The total losses for the receiver antenna can be calculated by adding all of the losses together, assuming that their values are in dB:

$$L_{ra} = L_{rr} + L_{r\text{pol}} + L_{r\text{foc}} + L_{r\text{point}} + L_{r\text{con}}$$

This total loss, as was the case in the transmitter section, is a direct attenuation of the signal: a 1-dB loss equals a 1-dB loss in the analysis.

1.4.2 Receiver antenna gain

The gain of the receiver antenna (in dB) is calculated in the same way as the transmitter antenna gain:

$$G_r = 10\log\left[n(\pi(D)/\lambda)^2\right]$$

where n is the efficiency factor <1, D is the diameter of the parabolic dish, and λ is the wavelength.

The antenna receiver gain is achieved by the size of the aperture and how much power it can collect from the scattered signal of the transmitter. The larger the aperture, the more surface area is available to receive the transmitted power.

The receiver antenna is not required to have the same antenna as the transmitter. The receiver can use an omnidirectional antenna and receive transmissions from a transmitter that uses a parabolic dish antenna, or the transmitter can be an omnidirectional antenna and the receiver can use a parabolic dish. However, if this is a direct line-of-sight system, the antennas should have the same polarization. For example, if a transmitter is using a vertically polarized antenna, the receiver should have a vertically polarized antenna. An exception to this rule is that if the system is using the ionosphere to bounce the signal for maximum range using lower frequencies, then the polarization can be reversed by the reflection off the ionosphere. For example, if a transmitter is using a horizontally polarized antenna, the optimal receiver antenna may be vertically polarized if the reflection off the ionosphere causes reversal of the polarization. The gain of the antenna is a direct gain in the communications link: a 1-dB gain equals a 1-dB improvement in the link.

1.4.3 Receiver line losses from the antenna to the LNA

The cable that connects the antenna to the first amplifier, which is designated as a LNA, is included in the total losses:

$$L_{rll} = \text{coaxial or waveguide line losses (in dB)}$$

The amplifier is referred to as a LNA because it is designed to have very low noise or noise figure (NF). This is important in setting the NF of the system, since it is determined mainly by the first amplifier in the receiver. A discussion and calculation of the NF of a system are provided later in this chapter. The important thing is that the NF increases the noise level on a one-for-one dB basis. The cable loss between the antenna and the LNA, as was the case in the transmitter section, is a direct attenuation of the signal: a 1-dB loss equals a 1-dB loss in the analysis. Therefore, as mentioned earlier, the cable length should be kept as short as possible, with the option of putting the LNA with the antenna assembly.

1.4.4 Receiver component losses

Any components between the antenna and LNA will reduce the signal-to-noise ratio (SNR) of the system. For example, often a limiter is placed in the line between the antenna and the LNA to protect the LNA from damage by high-power signals. This can be included in the NF of the receiver or viewed as a loss of the signal. Both methods are used in the industry, with the same end results, since the first method increases the noise and the second method decreases the signal level, producing the same SNR results for the receiver.

However, since this loss does not add noise to the system but only attenuates the signal, a more straightforward approach would be to treat it as a loss in signal

level and calculate the NF separately. The noise before and after the lossy devices is the same, since the temperature before and after the device is the same. Only the signal level is attenuated. The noise on the front end of the receiver before the LNA is equal to kTB, where k is the Boltzmann constant (1.38×10^{-23} J/K), T is the nominal temperature (290 K), and B is the bandwidth.

For a 1-Hz bandwidth,

$$kTB = 1.38 \times 10^{-23}\,\mathrm{J/K} \times 290\,\mathrm{K} \times 1\,\mathrm{Hz} = 4.002 \times 10^{-21}\,\mathrm{W} = 4.002 \times 10^{-18}\mathrm{mW}$$

Converting to dBm,

$$10\log(kTB) = 10\log(4.002 \times 10^{-18}) = -174\,\mathrm{dBm}$$

A convenient way to calculate the kTB noise is to use the aforementioned 1-Hz bandwidth number and simply take 10 log(B) and add to it. For example, for a 1-MHz bandwidth,

$$\begin{aligned}&-174\ \mathrm{dBm} + 10\ \log(1\ \mathrm{MHz})\ \mathrm{dB} = -174\ \mathrm{dBm} + 10\ \log(10^6\ \mathrm{Hz})\ \mathrm{dB}\\ &\quad = -174\ \mathrm{dBm} + 60\ \mathrm{dB} = -114\ \mathrm{dBm}\end{aligned}$$

The LNA amplifies the input signal and noise, and during this amplification process additional noise is present at the output, mainly due to the active transistors in the amplifier. This additional noise is referred to as the NF of the LNA and increases the overall noise floor, which reduces the SNR. Since this is a change in noise level, NF is in dB. The resultant noise floor is equal to $kTBF$, where F is the increase in the noise floor due to the LNA. The noise factor F is the increase in noise due to the amplifier, and the NF is the increase in noise converted to dB. The rest of the components of the receiver increase the $kTBF$ noise floor, but this contribution is relatively small. The amount of increased noise is dependent on how much loss vs gain there is after the LNA and if the noise floor starts approaching the kTB noise. Also, if the bandwidth becomes larger than the initial kTB bandwidth at the LNA, then the NF can degrade. Calculation of the actual receiver noise is discussed later. The important concept is that the LNA is the main contributor in establishing the noise for the receiver and must be carefully designed to maintain the best SNR for detection of the desired signal with the lowest NF. The kTB noise is a constant noise floor unless the temperature or bandwidth is changed. Therefore, the receiver component losses before the LNA are applied to the signal, thus reducing the SNR:

$$L_{r\mathrm{comp}} = \text{switches, circulators, limiters, and filters}$$

This loss, as was the case in the transmitter section, is a direct attenuation of the signal: a 1-dB loss equals a 1-dB loss in the link analysis.

The total losses between the receiver antenna and the LNA are therefore equal to

$$L_{tr} = L_{rll} + L_{r\mathrm{comp}}$$

1.4.5 Received signal power at the output to the LNA

The received signal level P_s (in dBm) at the output of the LNA is calculated as follows:

$$P_{s\ \mathrm{dBm}} = \mathrm{EIRP}_{\mathrm{dBm}} - A_{fs\ \mathrm{dB}} - L_{p\ \mathrm{dB}} - L_{\mathrm{multi\ dB}} - L_{ra\ \mathrm{dB}} + G_{r\ \mathrm{dB}} - L_{rll\ \mathrm{dB}} - L_{r\mathrm{comp\ dB}} + G_{\mathrm{LNA\ dB}}$$

where $\mathrm{EIRP}_{\mathrm{dBm}}$ is the effective radiated isotropic power, $A_{fs\ \mathrm{dB}}$ is the free-space attenuation, $L_{p\ \mathrm{dB}}$ is the propagation loss, $L_{\mathrm{multi\ dB}}$ is the multipath losses, $L_{ra\ \mathrm{dB}}$ is the total receiver antenna losses, $G_{r\ \mathrm{dB}}$ is the receiver antenna gain, $L_{rll\ \mathrm{dB}}$ is the coaxial or waveguide line losses, $L_{r\mathrm{comp\ dB}}$ is the switches, circulators, limiters, and filters, and $G_{\mathrm{LNA\ dB}}$ is the gain of the LNA.

Most often dBm is used as the standard method of specifying power.

The noise out of the LNA (in dBm) is equal to:

$$N_{\mathrm{LNA\ dBm}} = kTB_{\mathrm{dBm}} + NF_{\mathrm{dB}} + G_{\mathrm{LNA\ dB}}$$

Thus, a preliminary SNR can be calculated to obtain an estimate of the performance of a receiver by using the SNR at the output of the LNA. Since the SNR is a ratio of the signal and noise, and since both the signal power and the noise power are in dBm, a simple subtraction of the signal minus the noise will produce a SNR in dB as follows:

$$\mathrm{SNR}_{\mathrm{dB}} = P_{s\ \mathrm{dbm}} - N_{\mathrm{LNA\ dbm}}$$

where SNR is the signal-to-noise ratio (in dB), P_s is the power out of the LNA (in dBm), and N_{LNA} is the noise power out of the LNA (in dBm).

One point to note is that the gain of the LNA (G_{LNA}) is applied to both the signal level and the noise level. Therefore, this gain can be eliminated for the purposes of calculating the SNR at the output of the LNA.

$$\mathrm{SNR}_{\mathrm{dB}} = (P_{s\ \mathrm{dbm}} + G_{\mathrm{LNA\ dB}}) - (N_{\mathrm{LNA\ dbm}} + G_{\mathrm{LNA\ dB}}) = P_{s\ \mathrm{dbm}} - N_{\mathrm{LNA\ dbm}}$$

1.4.6 Receiver implementation loss

When implementing a receiver in hardware, there are several components that do not behave ideally. Losses associated with these devices degrade the SNR. One of the contributors to this loss is the phase noise or jitter of all of the oscillators and potential noise sources in the system, including the upconversion and downconversion oscillators, code and carrier tracking oscillators, phase-locked loops, match filters, and analog-to-digital converters. Another source of implementation loss is the detector process, including nonideal components and quantization errors. These losses directly affect the receiver's performance. Implementation losses (L_i) are included to account for the departure from the ideal design due to hardware implementation.

A ballpark figure for implementation losses for a typical receiver is about −3 dB. With more stable oscillators and close attention during receiver design, this can be reduced. This loss will vary from system to system and should be analyzed for each receiver system. This loss is a direct attenuation of the signal: a 1-dB loss equals a 1-dB loss in the link analysis.

1.4.7 Doppler effects on received signal

Doppler changes the frequency due to movement between the transmitter and the receiver that can cause frequency and phase errors in the detection of the desired incoming signals. When the receiver is moving away from the transmitter, the Doppler effect decreases the frequency and when the receiver is moving towards the transmitter the Doppler effect increases the frequency. An example using sound waves is shown in Figure 1.8. As a train is approaching or going towards a person, the sound of the train is at a higher frequency and when the train passes and begins to move away from the person, the sound of the train is at a lower frequency. This is the result of Doppler effects on sound waves. The same Doppler effects occur at RF frequencies and can cause distortion of the desired signal (Figure 1.9).

The equation for Doppler is as follows:

$$\Delta f = \frac{v_r}{c_o} \times f_o = \frac{v_r}{\lambda_o}$$

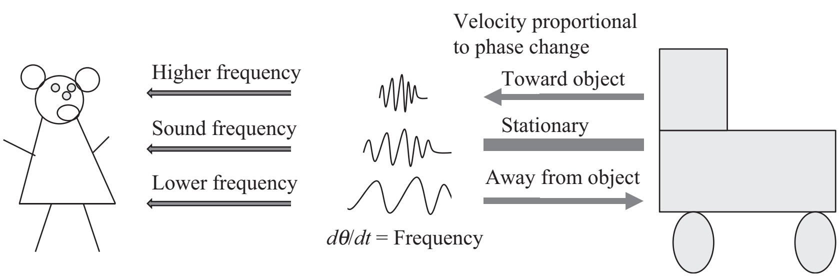


Figure 1.8 Sound waves exhibit Doppler effects

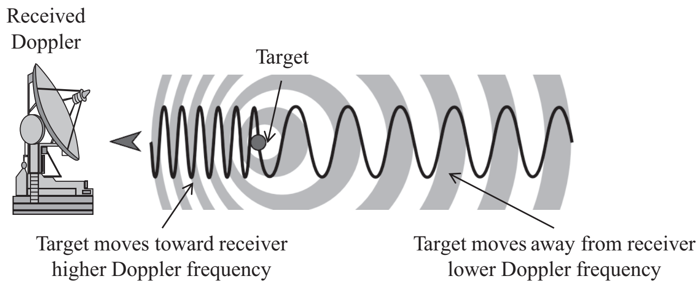


Figure 1.9 Doppler effects for moving targets

where Δf is the Doppler frequency, v_r is the radial velocity, f_o is the frequency, c_o is the speed of light, and λ_o is the wavelength of frequency.

Doppler affects the carrier frequency and phase and can distort the desired signal. The Doppler effects depends on the radial velocity between the transmitter and the receiver. This changes the phase and frequency and distorts the received signal.

Another example shows balls that are sent to the receiver as the transmitter is moving towards a receiver, the balls are received faster which is at a higher frequency (Figure 1.10). Doppler is determined by the radial velocity of the movement, where the transmitter is moving directly to the receiver (Figure 1.11). This provides the maximum detectable velocity of Doppler effects. If the velocity of the transmitter is perpendicular to the receiver, then the radial velocity is equal to zero and there is no Doppler effects. If the transmitter is moving at an angle to the receiver, then the radial velocity is calculated by taking the cosine of the angle multiplied by the velocity of the transmitter (Figure 1.12).

Since digital communications rely on accurate phase and frequency measurements, Doppler effects can cause errors in the detection and degrades the desired signal depending on the amount of Doppler.

1.4.8 Received power for establishing the signal-to-noise ratio of a system

The received power level from the transmitter to the receiver is shown in Figure 1.13. The transmitter power or EIRP is transmitted out of the transmitter antenna, through the channel with different losses, and finally into the receiver with the gain of the antenna and internal losses. The power at the receiver is the summation of all of the higher level blocks.

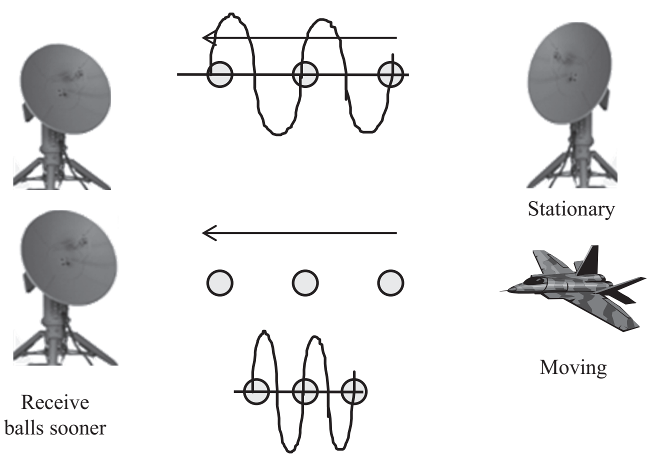


Figure 1.10 One-way Doppler effects for moving targets

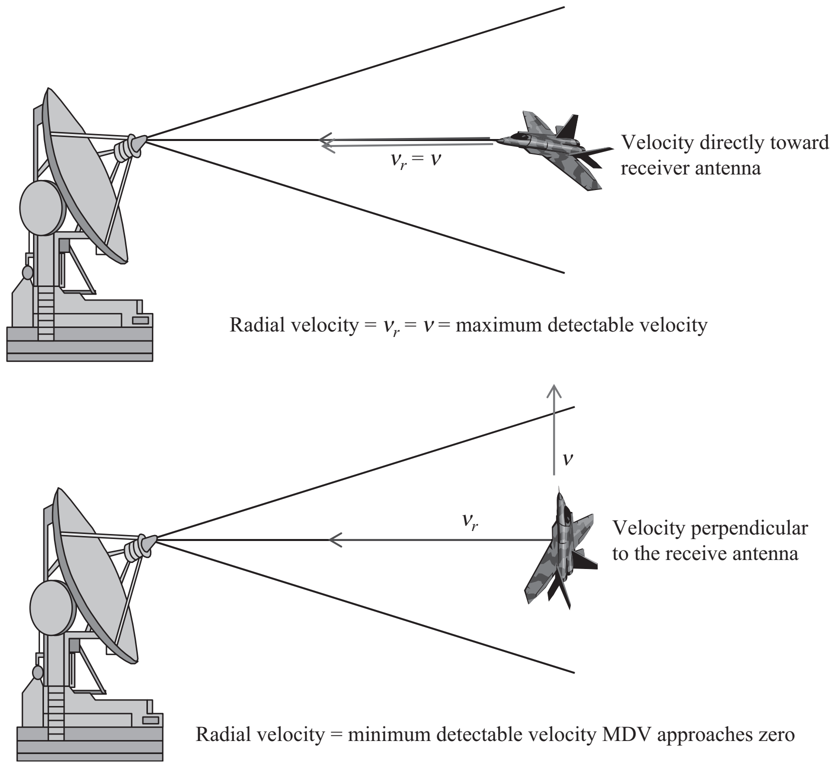


Figure 1.11 Radial velocity detectability for Doppler

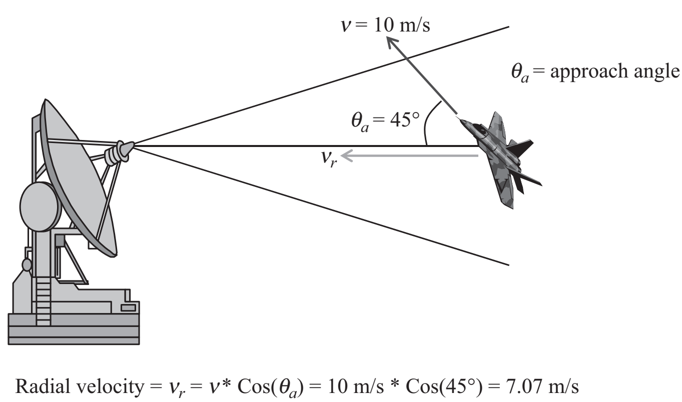
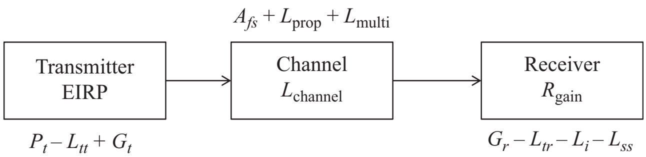


Figure 1.12 Radial velocity is a function of approach angle



Received signal power P_{rs}(dBm) = EIRP $- L_{channel} + R_{gain}$

$P_{rs}(\text{dBm}) = P_t - L_{tt} + G_t - A_{fs} - L_{prop} - L_{multi} + G_r - L_{tr} - L_i - L_{ss}$

Figure 1.13 Receiver signal power level

The detected power P_d or S (in dBm) that is used to calculate the final SNR for the analysis is:

$$P_{d\ \text{dBm}} = P_{s\ \text{dBm}} - L_{i\ \text{dB}} + G_{\text{receiver dB}} = S_{\text{dBm}}$$

where $P_{s\ \text{dBm}}$ is the power to the LNA, $L_{i\ \text{dB}}$ is the implementation losses, $G_{\text{receiver dB}}$ is the receiver gain, and $P_{d\ \text{dBm}} = S_{\ \text{dBm}}$, which is the detected power or signal level.

Carrying the receiver gain through the rest of the analysis is not necessary for SNR calculations since it applies to both the signal and the noise equally.

A typical example of analyzing the receiver power is shown below:

$f = 2.4$ GHz, Range = 10 mi = 16.1 km
PA output Pt = 60 dBm
Cable and circulator loss $L_{tt} = 3$ dB
Gain of the transmit antenna $G_t = 2.1$ dB
Free-space loss $A_{fs} = 20\log[(4\pi Rf)/c] = 124.2$ dB
Propagation loss $L_{prop} = 1$ dB
Multipath loss = 3 dB
Receiver Antenna Gain $G_r = 2.1$ dB
Cable and circulator loss $L_{tr} = 3$ dB
Implementation loss $L_i = 1$ dB
Spread spectrum loss $L_{ss} = 2$ dB
$P_{rs(\text{dBm})} = P_t - L_{tt} + G_t - A_{fs} - L_{prop} - L_{multi} + G_r - L_{tr} - L_i - L_{ss}$
$P_{rs(\text{dBm})} = 60\ \text{dBm} - 3\ \text{dB} + 2.1\ \text{dB} - 124.2\ \text{dB} - 1\ \text{dB} - 3\ \text{dB} + 2.1\ \text{dB} - 3\ \text{dB} - 1\ \text{dB} - 2\ \text{dB} = -73\ \text{dBm}$

1.4.9 Received noise power

The noise power (N) of the receiver is compared to the signal power (S) of the receiver to determine a power SNR. The noise power is initially specified using temperature and bandwidth. System-induced noise is added to this basic noise to establish the noise power for the system.

1.4.10 Noise figure

The standard noise equation for calculating the noise out of the receiver is:

$$N = kT_oBF$$

where T_o is the nominal temperature (290 K), F is the noise factor, k is the Boltzmann constant (1.38×10^{-23} J/K), and B is the bandwidth in Hz.

The noise factor is the increase in noise level that is caused by an active device generating noise that is greater than the kTB noise. For example, if the noise at the output of an active device increases the kTB noise by 4, then the noise factor would be 4. The NF is the noise factor (F) in dB. For example,

$$\begin{aligned} \text{Noise factor} &= F = 4 \\ \text{NF} &= F\ \text{dB} = 10\log(4) = 6\ \text{dB} \end{aligned}$$

If the NF is used, then the noise out of the receiver is in dBm and the equation is

$$N\ \text{dBm} = 10\log(kTB) + NF\ \text{dB}$$

Therefore, when solving for the noise of the receiver using the SNR, the equations become:

$$\begin{aligned} \text{Noise factor}\ F &= \text{SNR}_{\text{in}}/\text{SNR}_{\text{out}} \\ \text{NF(dB)} &= 10\log(\text{SNR}_{\text{in}}) - 10\log(\text{SNR}_{\text{out}}) = \text{SNR}_{\text{in dB}} - \text{SNR}_{\text{out dB}} \end{aligned}$$

The noise out of the receiver will be $kT_o\,BF$ plus any effects due to the difference in temperature between the sky temperature (at the antenna) and the nominal temperature (at the LNA), with F being the receiver noise factor:

$$N = kT_oBF + (kT_sB - kT_oB) = kT_oBF + kB(T_s - T_o) = kT_oBF_t$$

Solving for F_t

$$F_t = F + (T_s - T_o)/T_o$$

where T_s is the sky temperature and T_o is the nominal temperature (290 K).

If $T_s = T_o$, then the noise is $N = kT_oBF$. Since, generally T_s is less than T_o, the noise will be less than the standard kT_oBF and the noise factor is reduced by $(T_s - T_o)/T_o$.

The noise factor for the entire receiver is calculated using the Friis noise equation;

$$F_t = F_1 + [(F_2 \times \text{Losses}) - 1]/G_1 + [(F_3 \times \text{Losses}) - 1]/G_1G_2 \ldots$$

As more factors are added, they contribute less and less to the total noise factor due to the additional gains. The receiver noise factor is approximated by the LNA noise factor, F_1, assuming that the losses are not too great between stages. For the LNA, noise factor of F_1 and a gain of G_1, and one additional amplifier stage, noise factor F_2 with losses in between them, the total noise factor is calculated as follows

(Figure 1.14, Method 1).

$$F_t = F_1 + [(F_2 \times \text{Losses}) - 1]/G_1$$

Another method (Figure 1.14, Method 2) to utilize the Friis noise equation is to treat the losses as individual elements in the chain analysis as shown

$$F = F_1 + [F_2 - 1]/G_1 + [F_3 - 1]G_1G_2$$

Both of these methods produce the same results, as shown in Figure 1.14. The first method is the preferred one and will be used in the context of this publication.

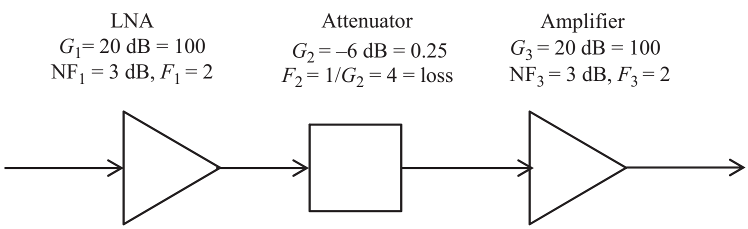
Note that the noise factor of the LNA, $F_1 = 2$, is approximately equal to the total noise factor for the receiver, $F_{\text{total}} = 2.07$. Therefore, the NF of the LNA (F_1) is the main contributor of the NF of the receiver. The exceptions would be if the losses are much greater in the chain or the bandwidth changes.

The losses before the LNA are not included in the NF calculation since they do not add noise to the system and are treated as losses to the signal. The reason for this is that the noise temperature is the same before and after the losses, so the kT noise is the same but the signal is attenuated. Therefore, the SNR is reduced on the output. There is no additional noise added to the system, but the NF has been defined as $\text{SNR}_{\text{in}} - \text{SNR}_{\text{out}}$ (in dB). This is true only if the bandwidths of the input and the output are the same. One major assumption is that the bandwidth is smaller as the signal progresses through the receiver. If the bandwidth increases, then the NF increases. This NF increase may be needed to be included in the analysis.

1.4.11 Received noise power at the detector

The noise power at the detector is computed by converting the parameters of the noise factors into dB to simplify the calculation. The gain of the receiver, G_{receiver},

$$F = F_1 + [F_2 - 1]/G_1 + [F_3 - 1]/(G_1G_2) + \ldots$$



$F_{\text{total}} = F_1 + [F_3 * F_2 - 1]/G_1 = 2 + [2*4 - 1]/100 = 2.07$

LNA $F_1 = 2$, basically sets NF of receiver, except for large losses between gain stages

Alternate method:

$F_{\text{total}} = F_1 + [F_2 - 1]/G_1 + [F_3 - 1]/(G_1G_2) = 2 + (4 - 1)/100 + (2 - 1)/(100*0.25) = 2.07$

Note: Do not use NF (in dB) in the Friis noise equation

Figure 1.14 System noise figure using the Friis noise equation

is included in this example:

$$N_{\text{dBm}} = kT_{o\ \text{dBm}} + B_{\text{dB}} + NF_{\text{LNA dB}} + G_{\text{receiver dB}}$$

Note that this assumes $T_s = T_o$ and that the line losses are included in the link losses and not in the NF.

Eliminating G_{receiver} for analysis purposes is optional since the gain is common to both the noise power and the signal power. However, it is often included to show the actual signal and noise levels through the receiver.

An example of the noise power parameters is the following:

- $N_{\text{ktbf}} = 10$ Log (kT_oBF)
- F is the noise factor, NF = 10 log(F)
- k = Boltzmann's constant (1.38×10^{-23} J/K)
- T_o = room temperature (290 K)
- B = bandwidth at the detector
- $N_{\text{ktb}} = -174$ dBm for 1 Hz bandwidth
- $N_{\text{ktb}} = -114$ dBm for 1 MHz bandwidth
- Example: Bandwidth = 10 MHz
- $N_{\text{ktb}} = -174$ dBm + 10 Log(10 MHz) = −174 dBm + 70 dB = −104 dBm
- $N_{\text{ktb}} = -114$ dBm + 10 Log(10) = −114 dBm + 10 dB = −104 dBm
- Given receiver NF = 3 dB
- $N_{\text{ktbf}} = -104$ dBm + 3 dB = −101 dBm
- Therefore, the SNR = −73 dBm – (−101 dBm) = 28 dB

1.4.12 Receiver bandwidth

The SNR is used for analog signals where the signal (S) is the signal power of the analog signal, and noise (N) is the amount of noise power in the required bandwidth for sending the analog signal. For digital modulation, square pulses or bits are used to send information from the transmitter to the receiver. The amount of energy in a single bit is denoted as E_b. The bit rate or the number of bits per second is denoted as R_b. The signal power for a digital modulation system is equal to the energy in a single bit times the bit rate:

$$S = E_b \times R_b$$

The noise power (N) is equal to the noise power spectral density (N_o), which is the amount of noise per Hz, or in other words, the amount of noise in a 1-Hz bandwidth, times the bandwidth (B) of the digital system:

$$N = N_o \times B$$

Therefore, the SNR for the digital system is equal to

$$\begin{aligned} S &= E_b \times R_b \\ N &= N_o \times B \\ \text{SNR} &= E_b/N_o \times R_b/B \end{aligned}$$

The bandwidth for a basic digital system is equal to the bit rate. In this case, the SNR is equal to:

$$\text{SNR} = E_b/N_o$$

For more complex digital communications, the bandwidth may not equal the bit rate, so the entire equation needs to be carried out.

For example, one method of digital modulation is to change the phase of a frequency in accordance with the digital signal, a digital "0" is 0° phase and a digital "1" is 180° phase. This is known as binary phase-shift keying (BPSK). (The types of digital modulation are discussed later in this book.) For BPSK, the bandwidth is equal to the bit rate:

$$\text{SNR} = E_b/N_o \times R_b/B$$
$$\text{SNR} = E_b/N_o$$

However, if a more complex digital modulation waveform is used that contains four phase states—0°, 90°, 180°, −90°—such as quadrature phase-shift keying (QPSK), then the bit rate is twice as fast as the bandwidth required because there are two bits of information for one phase shift. Further explanation is provided in Chapter 2. Consequently, the SNR is equal to:

$$\text{SNR} = E_b/N_o \times 2R_b/B$$
$$\text{SNR} = 2E_b/N_o$$

The bandwidth used in the BPSK example uses a bit rate equal to the bandwidth. This provides an approximation so that the SNR approximates E_b/N_o. E_b/N_o is used to find the probability of error from the curves shown in Chapter 6. The curves are generated using the mathematical error functions, *erf*(x).

For example, using BPSK modulation, the probability of error, P_e, is equal to:

$$P_e = 1 \times 10^{-8} \text{ for } E_b/N_o = 12 \text{ dB}$$

This equation gives the probability of bit error for an ideal system. If this is too low, then the E_b/N_o is increased. This can be accomplished by increasing the power out of the transmitter. This can also be accomplished by changing any of the link parameters discussed previously, such as by reducing the line loss. Either increasing E_b or decreasing N_o will increase the ratio.

For QPSK, since it has four phase states, it contains 2 bits of information for every phase state or symbol:

$$2^{\#\text{bits}} = \text{states}, 2^2 = 4$$
$$\text{Symbol rate of QPSK} = \text{Bit rate of BPSK}$$

Although it sends twice as many bits in the same bandwidth, it requires a higher SNR:

$$\text{SNR} = E_b/N_o \times 2\ BR/BW$$

If the symbol rate is reduced by 1/2 so that the bit rate is equal for both QPSK and BPSK, then both the bandwidth and noise are reduced by 1/2 and the probability of error is approximately the same.

1.4.13 Received E_b/N_o at the detector

The received E_b/N_o (in dB) is equal to the received SNR (in dB) for BPSK. For higher order modulations, the bandwidth will be larger than the bit rate, so the SNR (in dB) $= XE_b/N_o$, where X relates to the order of modulation.

E_b/N_o is the critical parameter for the measurement of digital system performance. Once this has been determined for a digital receiver, it is compared to the required E_b/N_o to determine if it is adequate for proper system operation for a given probability of error.

1.4.14 Receiver coding gain

Another way to improve the performance of a digital communication system is to use codes that have the ability to correct bits that are in error. This is called error correction. The most common error correction is known as forward error correction (FEC). This type of correction uses code to correct bit errors without any feedback to the transmitter. Coding will be discussed in more detail in subsequent chapters. With FEC, the coding gain is the gain achieved by implementing certain codes to correct errors, which is used to improve the bit error rate (BER), thus requiring a smaller E_b/N_o for a given probability of error. The coding gain, G_c depends on the coding scheme used. For example, a Reed–Solomon code rate of 7/8 gives a coding gain of approximately 4.5 dB at a BER of 10^{-8}. Note that coding requires a tradeoff, since adding coding either increases the bandwidth by 8/7 or the bit rate is decreased by 7/8.

1.4.15 Required E_b/N_o

The required E_b/N_o for the communications link is

$$E_b/N_o(\text{req})_{\text{dB}} = E_b/N_o(\text{uncoded})_{\text{dB}} - G_{c\ \text{dB}}$$

This is the minimum required $E_b/N_{o\ \text{dB}}$ to enable the transceiver to operate correctly. The parameters of the communications link need to be set to ensure that this minimum $E_b/N_{o\ \text{dB}}$ is present at the detector.

1.5 The link budget

The link budget is a method used to determine the necessary parameters for successful transmission of a signal from a transmitter to a receiver. The term "link" refers to linking or connecting the transmitter to the receiver, which is done by sending out RF waves through space (Figure 1.15). The term budget refers to the allocation of RF power, gains, and losses and tracks both the signal and the noise levels throughout the entire system, including the link between the transmitter and

the receiver. The main items that are included in the budget are the required power output level from the transmitter PA, the gains and losses throughout the system and link, and the SNR for reliable detection; the E_b/N_o to produce the desired BER; or the probability of detection and probability of false alarm at the receiver for pulse or radar systems. Therefore, when certain parameters are known or selected, the link budget allows the system designer to calculate unknown parameters.

Several of the link budget parameters are given or chosen during the process and the rest of the parameters are calculated. There are many variables and trade-offs in the design of a transceiver, and each one needs to be evaluated for each system design. For example, there are trade-offs between the power output required from the PA and the size of the antenna. The larger the antenna (producing more gain), the less power is required from the PA. However, the cost and size of the antenna may be too great for the given application. On the other hand, the cost and size of the PA increase as the power output increases, which may be the limiting factor. If the power output requirement is large enough, a solid-state amplifier may not be adequate, and therefore a traveling-wave tube amplifier (TWTA) may be needed. The TWTA requires a special high-voltage power supply, which generally increases size and cost. By making these kinds of trade-off studies, an optimum data link solution can be designed for a specific application.

Before starting the link budget, all fixed or specified information concerning the transceiver needs to be examined to determine which parameters to calculate in the link budget. These concessions need to be evaluated before the link budget is performed and then must be reevaluated to ensure that the right decisions have been made. The parameters for a link budget are described previously in this chapter.

Proper transceiver design is critical in the cost and performance of a data link. To provide the optimal design for the transceiver, a link budget is used to allocate the gains and losses in the link and to perform trade-offs of various parts of the system. The link budget also uses the required SNR or the ratio of bit energy to noise spectral density (E_b/N_o) for a given probability of error. These required levels are derived by using probability of error curves given a certain type of modulation. Probability of error curves are discussed in Chapter 6. Generally, since there are both known and unknown variances in the link budget, a link budget will provide an additional SNR or E_b/N_o, which is referred to as the link margin. The link

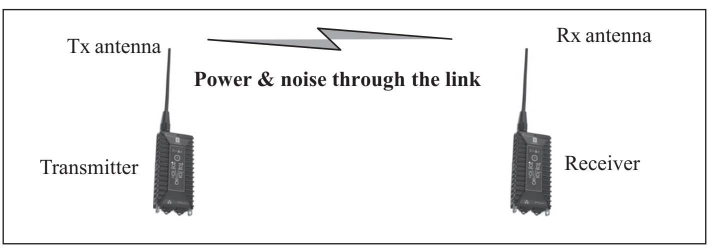


Figure 1.15 Link connects the transmitter to the receiver for link budget analysis

margin is equal to:

$$\text{Link margin for analog systems} = \text{SNR}(\text{calculated})_{\text{dB}} - \text{SNR}(\text{required})_{\text{dB}}$$
$$\text{Link margin for digital systems} = E_b/N_o(\text{calculated})_{\text{dB}} - E_b/N_o(\text{required})_{\text{dB}}$$

This link margin is used to provide a margin of error in the analysis, hardware implementation, and other factors that can affect the desired performance.

1.5.1 Spread spectrum systems

Spread spectrum is a technique for using more bandwidth than is required to send digital data. This is generally accomplished by using a faster code sequence and combining it with the digital data sequence before modulation. Spread spectrum is used mainly to mitigate jamming signals. To complete the link analysis for a spread spectrum system, spreading losses need to be included in the link budget.

If spread spectrum is used, then the signal level is modified by the spreading losses associated with nonideal spreading and despreading techniques, which reduce the received signal power (sometimes referred to as match filter loss). This is an additional loss separate from the implementation loss, since it deals with the spreading and despreading of a spread spectrum system. Also, since many digital systems are not spread spectrum systems, the loss is kept separate for clarity. The spread spectrum loss is generally around 1–2 dB and varies from system to system. Note that this is not included in a nonspread spectrum system. The losses are specified as follows:

$$L_{ss} = \text{spread spectrum loss}\ (1 - 2\ \text{dB})$$

The value of the spread spectrum loss is dependent on the type of spread spectrum used and the receiver design for spreading and despreading. This loss is a direct attenuation of the signal: a 1-dB loss equals a 1-dB loss in the link analysis.

1.5.2 Process gain

One of the main reasons for using spread spectrum is the ability of this type of system to reject other signals or jammers. This ability is called process gain (G_p). The process gain minus spreading losses provides the ideal jamming margin for the receiver. The jamming margin is the amount of extra power, referenced to the desired signal, which the jammer must transmit to jam the receiver:

$$J_{m\ \text{dB}} = G_{p\ \text{dB}} - L_{ss\ \text{dB}}$$

1.5.3 Received power for establishing the signal-to-noise ratio for a spread spectrum system

The detected power, P_d or S (in dBm), that is used to calculate the final SNR for the analysis of spread spectrum systems includes the spreading loss, L_{ss}:

$$P_{d\ \text{dBm}} = P_{s\ \text{dBm}} - L_{i\ \text{dB}} - L_{ss\ \text{dB}} + G_{\text{receiver dB}} = S_{\text{dBm}}$$

where $P_{s\ \mathrm{dBm}}$ is the power to the LNA, $L_{ss\ \mathrm{dB}}$ is the spreading losses, $G_{\mathrm{receiver\ dB}}$ is the receiver gain.

The receiver gain is applied to both the signal and the noise and is canceled out during the SNR calculation. Again, carrying the receiver gain through the rest of the analysis is optional.

1.5.4 Link budget example

A typical link budget spreadsheet used for calculating the link budget is shown in Table 1.3. This link budget tracks the power level and the noise level side by side so that the SNR can be calculated anywhere in the receiver. The altitude of an aircraft or satellite and its angle incident to the earth are used to compute the slant range, and the atmospheric loss is analyzed at the incident angle. The bit rate plus any additional bits for FEC determines the final noise.

A block diagram of a typical transmitter and receiver is shown in Figure 1.16. This block diagram refers to the link analysis that was performed in accordance with the typical link budget analysis (Table 1.3).

Table 1.3 Typical link budget analysis

Link Budget Analysis:			
	Slant Rng(km)	**Freq.(GHz)**	**Power(W)**
Enter Constants…	92.65	3	1
Enter Inputs………	**Inputs**	**Power Levels**	
Transmitter	Gain/Loss (dB	Sig.(dBm)	Noise(dBm)
Tran.Pwr(dBm)=		30	
Trans. line loss =	-1	29	
Other(switches)	-2	27	
Trans Ant Gain =	10	37	
Ant. Losses=	-2	35	
ERP		35	
Channel			
Free Space Loss =	-141.32	-106.32	
Rain Loss =	-2.00	-108.32	
Cloud Loss =	-1.00	-109.32	
Atm loss(etc) =	-0.50	-109.82	
Multipath Loss=	-2.00	-111.82	
Receiver			
RF BW(MHz)	100.00		-94.00
Ant. losses =	-2.00	-113.82	
Rx Ant Gain =	10.00	-103.82	
Ant. ohmic loss	-2.00	-105.82	
Other(switches)	-1.00	-106.82	
Rec. Line loss =	-2.00	-108.82	
LNA Noise Fig. =	3.00		-91.00
LNA Gain =	25.00	-83.82	-66
NF Deviation =	0.10		-65.90
LNA levels =		-83.82	-65.90
Receiver Gain =	60.00	-23.82	-5.90
Imp. Loss	-4.00	-27.82	
Despread BW	10.00		-15.9
PG 10log# of bits	30.00	2.18	
Spreading Loss =	-2.00	0.18	
Match Filt. loss			
Res. Doppler			
Det. Peak offset			
Det. Levels =		0.18	-15.9
S/N =		**16.08**	
Req. Eb/No			
Req. Eb/No		12	Pe=10exp-8
Coding Gain =	4.00	8.00	
Eb/No Margin =		8.08	

Conversions:						
Slant Rng:	Enter Range		Kilometers			
	2	mi	3.218			
	50	nmi	92.65			
	5280	ft	1.61			
Trans. Pwr:	Enter Power		Pwr(watts)			
	50	dBm	100			
Trans. Ant Gain:		Enter	Gain (dB)	Parabolic Dish		
Diameter (inches) =		24	22.63			
1/2 pwr beamwidth=		11	23.01	Bm Meth		
n= (efficiency) =		0.5				
Rx Ant. Gain:		Enter	Gain (dB)	Parabolic Dish		
Diameter(inches) =		24	22.63			
1/2 pwr beamwidth=		11	23.01	Bm Meth		
n = (efficiency) =		0.5				
Calculated Constants:			**Prob.of Err**	**Eb/No(dB**	**Eb/No**	**Pe**
Lambda =	0.1	meters	Coh.PSK	10.53	11.29796	1.00E-06
Lambda =	3.94	inches	Coh.FSK	13.54	22.59436	1.00E-06
Free Space Lo	-141.32	dB	Non.(DPSK)	11.18	13.122	1.00E-06
Boltzman's =	1.4E-23		Non.FSK	14.19	26.24219	1.00E-06
			Coh.QPSK/M	10.53	11.29796	1.00E-06
Receiver Noise Figure:		8.00	dB	6.31	Factor	
Losses, Post Preamp:		10.00	dB	10.00		
LNA Gain:		25.00	dB	316.23		
LNA NF:		3.00	dB	2.00		
NF Final:		3.10	dB	2.04		
NF Deviation:		0.10				
Ant. losses		Transm	Receive			
Lpointing (misalign)		0.5	0.5			
Lpolarization =		0.5	0.5			
Lradome loss =		0	0			
Lconscan(xover)=		0	0			
Lfoc(refract, L-angle)=		1	1			

Enter the actual bandwidths in MHz in the column even though they are not gains and losses. The noise power is adjusted accordingly.

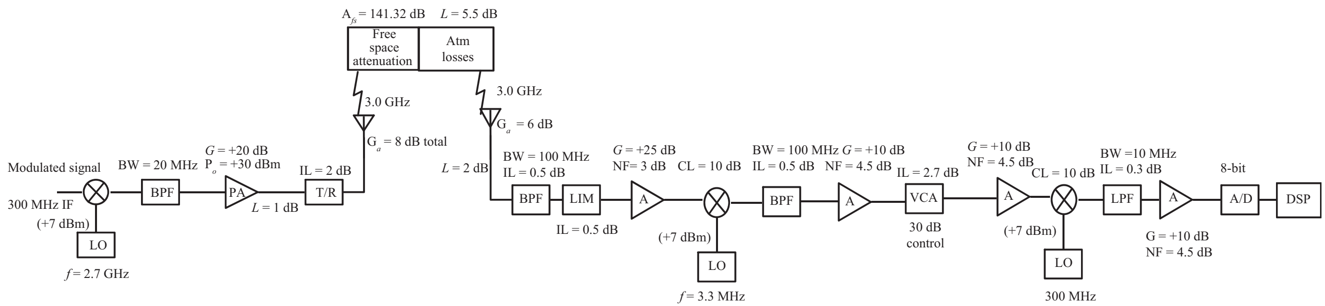


Figure 1.16 Data-link system for link budget analysis

1.5.5 Benefits of using a link budget

There are two basic benefits of using a link budget. They are:

1. Analysis to ensure the link works
   - (i) Calculates the signal and noise levels through the system
   - (ii) Determines if the E_b/N_o received is sufficient for the required P_e
   - (iii) Calculates link margin for anomalies and additional potential errors
   - (iv) Estimates of the path loss between the T_x and R_x
   - (v) Determines maximum range with specified P_e
   - (vi) Assists in the design of the hardware
   - (vii) Determines total required gain of the receiver
2. Optimization and trade-offs of the hardware
   - (i) Reduce the loss between the PA and LNA
   - (ii) Improve the LNA design to reduce the noise level
   - (iii) Reduce the losses between amplifier stages in the receiver—Friis
   - (iv) Size of antenna vs power output of PA

1.6 Summary

The transceiver design is applicable to all wireless communication systems. The link budget provides a means to design a transceiver and to perform the necessary analysis to ensure that the design is optimal for the given set of requirements. The link budget provides a way to easily make system trade-offs and to ensure that the transceiver operates properly over the required range. Once known or specified parameters are entered, the link budget is used to solve for the unknown parameters. If there is more than one unknown parameter, then the trade-offs need to be considered. Discussion of the differences between dB and dBm and their usage, E_b/N_o vs S/N, Doppler effects to the link margin, and NF calculations using the Friis noise equation are discussed in this chapter. The link budget is continually reworked, including the trade-offs, to obtain the best system design for the link. Generally, a spreadsheet is used for ease of changing the values of the parameters and monitoring the effects of the change (Table 1.3).

1.7 Problems

1. Show that a specification of 0 dBm $\pm$ 2 dBm = ? dBm is an impossible statement and write a correct statement for it.
2. What assumption is needed for S/N = E_b/N_o?
3. Solve for the total signal level through the receiver in Figure 1.P2 using milliwatts to obtain the answer, then solve for the output in dBm for the above receiver. Compare your answers. Note: Try the dB method without using a calculator.
4. An oscillator generates a center frequency of 1 GHz at 1 mW. The largest spurious response or unwanted signal is at a power level of 0.01 mW. How

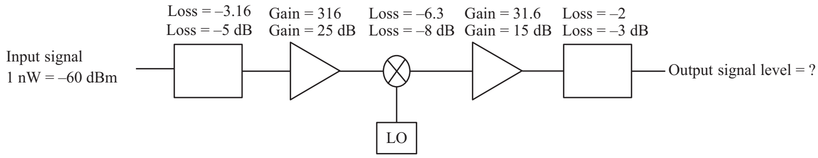


Figure 1.P2 A receiver diagram for calculating power output

would you specify the spurious response in dBc? What would the power be in milliwatts for a spur at −40 dBc?

5. A system has been operational for the past 3 years. A need arose to place a limiter in the path between the antenna and the LNA to avoid overload from a newly built high-power transmitter. The limiter has 1.5 dB of loss. How will this affect the ability to receive low-level signals? What can you do to overcome the loss?
6. What is the diameter of a parabolic antenna operating at 5 GHz, at an efficiency of 0.5 and a gain of 30 dBi?
7. What is the free-space attenuation for the system operating at 2.4 GHz with a range of 10 nmi?
8. Calculate the Doppler frequency given that the radial velocity is 100 km/h and a carrier frequency of 2.4 GHz?
9. Given that a 5.8-GHz transmitter is moving at 10 m/s at an angle of 45° to the receiver, what is the Doppler frequency?
10. What is the noise level out of the LNA given that the bandwidth is 10 MHz and the LNA NF is 3 dB with $T_o = T_s$ at room temperature?
11. What is the NF of the receiver, given that the LNA has an NF of 3 dB, a gain of 20 dB, a second amplifier after the LNA with a gain of 20 dB and an NF of 6 dB, and there is a loss of 5 dB between the amplifiers?
12. What is the purpose of a link budget?

Further reading

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Chapter 2
The transmitter

The transmitter is responsible for formatting, encoding, modulating, and upconverting the data communicated over the link using the required power output according to the link budget analysis. The transmitter section is also responsible for spreading the signal using various spread spectrum techniques. Several digital modulation waveforms are discussed in this chapter. The primary types of digital modulation using direct sequence methods to phase modulate a carrier are detailed, including diagrams and possible design solutions. A block diagram showing the basic components of a typical transmitter is shown in Figure 2.1.

2.1 Basic functions of the transmitter

The basic functions of the transmitter are the following:

- Transmitter antenna: provides an antenna to deliver power into space toward the receiver.
- Transmit/receive (T/R) device: T/R switch, circulator, diplexer, or duplexer allows the same antenna to be used for both the transmitter and the receiver.
- Radio frequency (RF) power amplifier (PA): provides an RF PA, which amplifies the signal power to an antenna.
- Upconverter: utilizes upconversion to select a carrier for the modulated signal that provides a frequency of operation in order to use a practical, smaller antenna. This upconverter could contain multiple upconverters as necessary to reach the final RF for transmission.
- Modulator: provides data modulation, which puts data on the carrier frequency. This can be accomplished using various types of modulation techniques, including spread spectrum and coding.

2.1.1 Transmit antenna

The antenna receives the RF signal from the PA through the T/R device and translates this RF power into electromagnetic waves so that the signal can be propagated through space. The antenna is dependent on the RF frequency selected by the upconverter according to the specified operational requirements. The proper design of the antenna ensures that the maximum signal power is sent out in the direction of the receiver or in the required service volume. The design of the

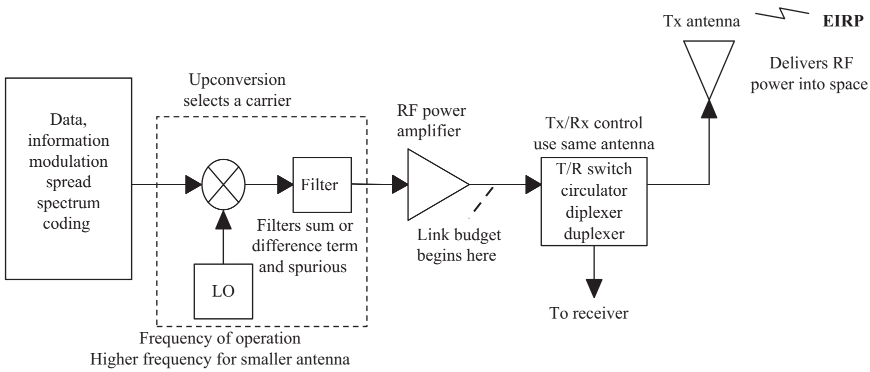


Figure 2.1 Block diagram of a basic transmitter

antenna is a critical parameter in the link budget and must be optimized to produce the best performance for a given size and cost.

The gain of the antenna improves the link budget on a one-for-one basis, that is, for every dB of gain, the antenna exhibits, the link is improved by a dB. Therefore, careful design of the antenna can reduce the power output required from the PA, which reduces the cost and thermal dissipation of the system. The frequency, amount of gain required, and size are factors in determining the type of antenna to use. Parabolic dishes are frequently used at microwave frequencies. The gain of the parabolic dish was calculated in Chapter 1. The antenna provides gain in the direction of the beam to reduce power requirements or increase range. The gain of the antenna is usually expressed in dBi, which is the gain in dB referenced to what an ideal isotropic radiator antenna would produce. In other words, this is the amount of amplifier gain that would be delivered to an isotropic radiator antenna to transmit the same amount of power in the same direction as a directional antenna. Even a vertical dipole antenna provides gain in the direction of the receiver, which is typically 2.14 dBi.

2.1.2 Transmit/receive device

To reduce the number of antennas for a transceiver system, which ultimately reduces size and cost, the same antenna is generally used for both the transmitter and the receiver at each location. If the same antenna is going to be used for the system, a T/R device is required to prevent the transmitter from interfering with or damaging the receiver. This device can be a duplexer or diplexer, T/R switch, circulator, or some combination of these devices. The method chosen must provide the necessary isolation between the transmitter circuitry and the receiver circuitry to prevent damage to the receiver during transmission. The transmitted signal is passed through this device and out to the antenna with adequate isolation from the receiver. The received signal is passed through this device from the antenna to the receiver with adequate isolation from the transmitter.

Duplexers and diplexers are commonly used to provide the isolation between the transmit signal and the receiver, but they are often misunderstood. A duplexer is a device that provides a passband response to the transmitted signal and a notch filter at the receiver frequency. Therefore, the transmitted output is passed through the duplexer to the antenna, but the transmitted signal at the receiver frequency is attenuated by the notch filter. Generally, the receiver frequency and the transmitter frequency are in the same band but separated far enough in frequency so that the bandpass filter and the notch filter can be separated. A diplexer is basically two bandpass filters, one tuned to the transmitter frequency band and the other tuned to the receiver frequency band. Diplexers are usually used to split multiple frequencies, such as two RF bands in the receiver, so that the same antenna can be used to receive two different frequency bands. For example, in the amateur radio bands, a single antenna can be used to receive the 2-m band and the 70-cm band because a diplexer can separate these bands using two bandpass filters in the receiver.

2.1.3 RF power amplifier

The RF PA is used to provide the necessary power output from the transmitter to satisfy the link budget for the receiver. The general classifications for PAs are as follows:

- Class A: Power is transmitted all of the time; used for linear operation.
- Class A/B: Power is transmitted somewhere in between Classes A and B. This is a trade-off between the linear operation of a Class A amplifier, and the efficiency of a Class B amplifier.
- Class B: The conduction angle for Class B is 180° or half the time. A classic implementation of Class B amplifiers for a linear system is push–pull, that is, one stage is on and the other stage is off. The only drawback of this configuration is crossover distortion when transitioning from one device to the other. Class B amplifiers are more efficient than Class A and A/B amplifiers, but they are more nonlinear.
- Class C: Power is transmitted less than half of the time. This amplifier class is more efficient than the previous classes, but it is more nonlinear. Class C amplifiers may present a potential problem for some types of digital modulation techniques.

PAs can be classified in many other ways depending on their design, function, and application, but the aforementioned list outlines the basic PAs used for communications.

To generate higher power outputs, PAs can be combined (Figure 2.2). The input is divided into multiple signal paths that are amplified by separate PAs and then coherently combined to generate higher power outputs. The phase/delay of each of the paths should be matched properly before they are combined, and a quadrature 90° hybrid can be used to reduce the reflected power and provide a better match. In addition, feedback techniques have been used to provide better matching characteristics for the combining.

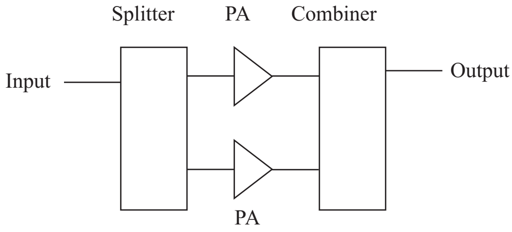


Figure 2.2 Combining power amplifiers for high power output

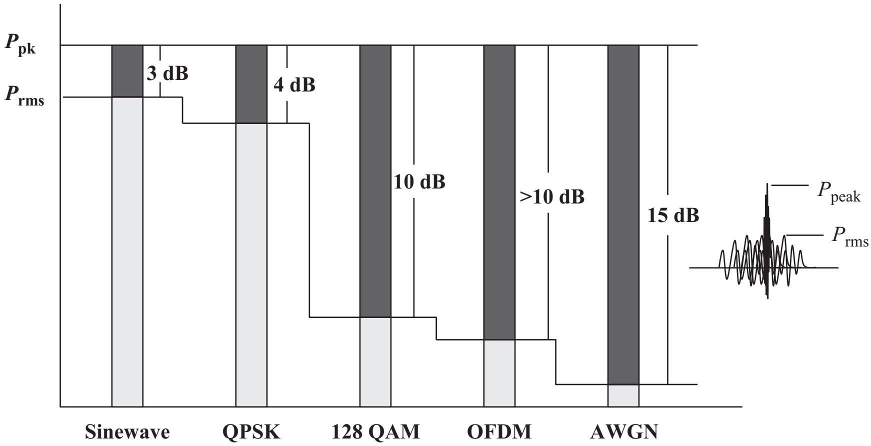


Figure 2.3 Crest factors for different signals

2.1.4 Crest factor

An objective for most transmitters is to obtain the highest power out of an amplifier without distortion due to saturation. To accomplish this, the waveform needs to be evaluated with respect to its crest factor (CF). The CF is equal to the required peak-to-average ratio or peak-to-average power ratio (PAPR) for each of the waveforms (Figure 2.3). This ratio is equal to the peak power divided by the average, or RMS power:

$$\mathrm{CF} = \mathrm{PAPR} = P_{\mathrm{peak}}/P_{\mathrm{rms}}$$
$$\mathrm{CF_{dB}} = \mathrm{PAPR_{dB}} = P_{\mathrm{peak\ dBm}} - P_{\mathrm{rms\ dBm}}$$

The requirement specifies that the transmitter needs to transmit the peak power without saturation. This causes the average power to be less so that the system requires a larger PA for the same performance. For example, a sinewave has a CF of 3 dB compared with an orthogonal frequency-division multiplexing (OFDM) waveform that has a CF of $>$10 dB. In order to transmit an OFDM waveform, the transmitter would have to have an additional 7 dB or greater power output requirement compared with a sinewave signal.

The CF for noise is high since some large noise spikes are well above average noise power output (Figure 2.3).

2.1.5 Upconverter

The information to be transmitted is made up of low frequencies, such as voice or a digital data stream. The low frequencies are mixed or upconverted to higher frequencies before the signal is radiated out of the antenna. One reason for using higher frequencies for radiation is the size of the antenna. The lower the frequency, the longer the antenna is required. For example, many antenna designs require antenna lengths of $\lambda/2$. The antenna length is calculated as follows:

$$\lambda = c/f$$

where λ is the wavelength, c is the speed of light $= 3 \times 10^8$ m/s, and f is the frequency.

Given:

$$\begin{aligned} f &= 10\ \text{MHz} \\ \lambda/2 &= (c/f)/2 = (3 \times 10^8/10\ \text{MHz})/2 = 15\ \text{m} = \text{length of the antenna.} \end{aligned}$$

This is much too large to use for many practical applications. If the frequency is upconverted to a higher frequency, then the antenna can be small as shown:

$$\begin{aligned} f &= 2.4\ \text{GHz} \\ \lambda/2 &= (c/f)/2 = (3 \times 10^8/2.4\ \text{GHz})/2 = 0.0625\ \text{m} \\ &= 6.25\ \text{in} = \text{length of the antenna.} \end{aligned}$$

This length of antenna is much more practical for multiple applications.

Some methods can be used to shorten the physical length while maintaining the required electrical length but not to a great extent, so the frequency still needs to be higher for a shorter antenna.

Another reason for using higher frequencies is band congestion. The lower usable frequency bands are overcrowded. Operating frequencies continue to increase in hopes that these bands will be less crowded. Regardless of the reason, the main function of the transmitter is to upconvert the signal to a usable radiating frequency.

A local oscillator (LO) is used to translate the lower frequency band to a higher frequency band for transmission. For example, a single lower frequency is mixed or multiplied by a higher frequency, which results in the sum and difference of the two frequencies as shown:

$$\begin{aligned} \cos(\omega_c t + \varphi_c)\cos(\omega_l t + \varphi_l) = 1/2\cos[(\omega_c - \omega_l)t + \varphi_c - \varphi_l] \\ + 1/2[\cos(\omega_c + \omega_l)t + \varphi_c + \varphi_l] \end{aligned}$$

where ω_c is the carrier or higher frequency, φ_c is the phase of the carrier, ω_l is the lower frequency containing the data, φ_l is the phase of the lower frequency, and t is the time.

A simpler form showing the concept of upconversion to a higher frequency is as follows:

$$f_c \times f_m = (f_c + f_m) + (f_c - f_m) = 2.4\text{ GHz} \times 10\text{ MHz} = (2.4\text{ GHz} + 10\text{ MHz}) + (2.4\text{ GHz} - 10\text{ MHz}) = 2.41\text{ GHz} + 2.39\text{ GHz}$$

where f_c is the high frequency carrier = 2.4 GHz and f_m is the low modulating frequency = 10 MHz or 0.010 GHz.

A diagram of the results of this mixing process is shown in Figure 2.4.

This is an ideal case that ignores any of the harmonics of the two input frequencies and only uses one low frequency (usually the data include a band of frequencies). A bandpass filter is used at the output to select either the sum term or the difference term. The sum term is selected in this example. A bandpass filter is centered at $\cos[(\omega_c + \omega_l)t + \varphi_c + \varphi_l]$ with a roll-off sharp enough to virtually eliminate the difference term $\cos[(\omega_c - \omega_l)t + \varphi_c - \varphi_l]$, with the following result:

$$1/2\cos[(\omega_c + \omega_l)t + \varphi_s]\text{or in simple form } (f_c + f_m)$$

where φ_s is the sum of the phases, $\omega_c + \omega_l$ is the sum of the frequencies, f_c is the high frequency carrier, and f_m is the low modulating frequency.

2.1.6 Sum and difference frequencies generated in the upconversion process

An LO is used to upconvert lower frequencies to higher frequencies for transmitting out the antenna. During this upconversion process, two frequency bands are produced in the ideal case: the sum of the LO frequency and the input frequencies, and the difference of the LO frequency and the input frequencies. One of these bands of frequencies must be filtered out to recover the signal at the receiver. In addition, since the same data are contained in both frequency bands, more power is required to transmit both sum and difference channels, and there is more chance of

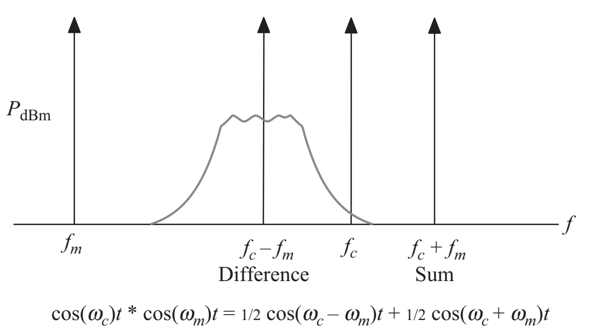


Figure 2.4 Ideal frequency spectrum of a CW modulating signal mixed with a CW carrier

generating interference to other uses by transmitting both bands. If both bands are allowed to be transmitted, there is a possibility that the phases will align at the receiver, which will cancel the desired signal. For example, suppose that the signal to be transmitted is defined as $A\cos(\omega_s)t$, and the carrier frequency is defined as $B\cos(\omega_c)t$. The output of the upconversion process is found by multiplying the two signals:

$$A\cos(\omega_s)t \times B\cos(\omega_c)t = AB/2\cos(\omega_s + \omega_c)t + AB/2\cos(\omega_s - \omega_c)t$$

Generally, one of these terms is filtered out. If both of these terms are allowed to be transmitted, a worst case phase situation at the receiver would be when the LO or carrier frequency ω_c is 90° out of phase $[C\cos(\omega_c t + 90) = C\sin(\omega_c)t]$ with the incoming signal:

$$\begin{aligned}
&[AB/2\cos(\omega_s + \omega_c)t + AB/2\cos(\omega_s - \omega_c)t][C\sin(\omega_c)t] \\
&\quad = ABC/4[\sin(\omega_s + \omega_c + \omega_c)t + \sin(\omega_s - \omega_c + \omega_c)t] \\
&\quad - ABC/4[\sin(-\omega_s + \omega_c - \omega_c)t + \sin(\omega_s - \omega_c - \omega_c)t] \\
&\quad = ABC/4[\sin(\omega_s + 2\omega_c)t + \sin(\omega_s)t] - ABC/4[\sin(\omega_s)t + \sin(\omega_s - 2\omega_c)t] \\
&\quad = ABC/4[\sin(\omega_s + 2\omega_c)t - \sin(\omega_s - 2\omega_c)t]
\end{aligned}$$

The results show that only the sum of $\omega_s + 2\omega_c$ and the difference of $\omega_s - 2\omega_c$ frequencies were detected by the receiver. The frequency ω_s was not retrieved, and the output was degraded substantially. This demonstrates the requirement to filter out one of the unwanted sidebands during transmission.

If the sum term is filtered out at the transmitter, and using the worst-case phase situation at the receiver as before, yields:

$$\begin{aligned}
&[AB/2\cos(\omega_s - \omega_c)t][C\sin(\omega_c)t] \\
&\quad = ABC/4[\sin(\omega_s - \omega_c + \omega_c)t] - ABC/4[\sin(\omega_s - \omega_c - \omega_c)t] \\
&\qquad = ABC/4[\sin(\omega_s)t - ABC/4[\sin(\omega_s - 2\omega_c)t]
\end{aligned}$$

The results of filtering the sum band in this worst-case scenario demonstrates that both the signal waveform ω_s and the difference of $\omega_s - 2\omega_c$ are retrieved. A low-pass filter will eliminate the second term, since the carrier frequency is generally much higher than the signal frequency, giving only the desired signal output of $ABC/4[\sin(\omega_s)t]$.

Another way to eliminate the unwanted band in the transmitter is to use an image-reject mixer. The image-reject mixer quadrature upconverts the signal into two paths: an in-phase path called the I-channel, and a path that is in quadrature or −90° out of phase called the Q-channel:

$$\begin{aligned}
\text{I channel output} &= \cos\omega_s t \times \cos\omega_c t = 1/2\cos(\omega_s - \omega_c)t \\
&\quad + 1/2\cos(\omega_s + \omega_c)t \\
\text{Q channel output} &= \cos\omega_s t \times \cos(\omega_c t - 90) \\
&= 1/2\cos[(\omega_s - \omega_c)t + 90] \\
&\quad + 1/2\cos[(\omega_s + \omega_c)t - 90]
\end{aligned}$$

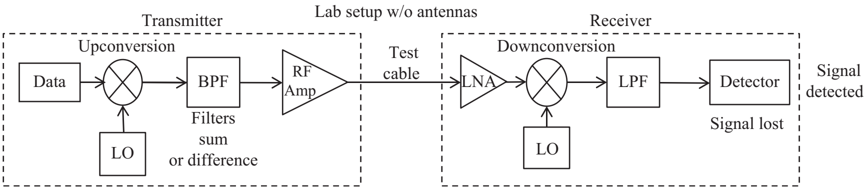


Figure 2.5 Lab example of attenuation of the desired signal when filters are not used

The I-channel output is then phase shifted by $-90°$, and the I and Q channels are then summed together, eliminating the difference band as follows:

$$\begin{aligned}
&\text{I-channel phase shifted} = 1/2\cos[(\omega_s - \omega_c)t - 90] + 1/2\cos[(\omega_s + \omega_c)t - 90] \\
&\text{I} + \text{Q} = 1/2\cos[(\omega_s - \omega_c)t - 90) + 1/2\cos[(\omega_s + \omega_c)t - 90] \\
&\qquad + 1/2\cos[(\omega_s - \omega_c)t + 90] + 1/2\cos[(\omega_s + \omega_c)t - 90] \\
&\quad = \cos[(\omega_s + \omega_c)t - 90]
\end{aligned}$$

Therefore, only the sum of the frequency band is transmitted, which eliminates the problem on the receiver side.

An example of this idea that can happen in a lab environment testing a transmitter and a receiver without filtering or antennas connected to the setup is shown in Figure 2.5. Without the filters between this setup as shown, there is a strong possibility that the desired signal will be attenuated depending on the phase drifts of the oscillators. The setup does not require antennas, but it does require filtering of either the sum or difference frequency to prevent problems with the detection.

2.2 Voltage standing wave ratio

The voltage standing wave ratio (VSWR) is caused by a mismatch in the impedance between devices. It determines how much power is reflected back from the load that is not matched to the source. Typically, this is specified using a 50-Ω system; the VSWR is the measurement used to indicate the amount of impedance mismatch. The amount of power that is reflected to the source is a loss in the signal level, which is a 1 dB, for 1-dB loss in the link analysis.

In the link analysis, the PA is chosen along with the gain of the antenna to meet the slant range of the data link. The PA is generally a high-cost item in the transceiver, and therefore, careful transceiver design to minimize losses in the link analysis will reduce the power requirement of the PA and also the overall cost of the transceiver. Therefore, designing the PA and the antenna to obtain the best impedance match or the lowest VSWR will help to minimize the loss.

The VSWR defines a standing wave that is generated on the cable between the PA and the antenna. The ratio of the minimum and maximum voltages measured along this line is the VSWR (Figure 2.6). For example, a VSWR of 2:1 defines a mismatch where the maximum voltage is twice the minimum voltage on the line.

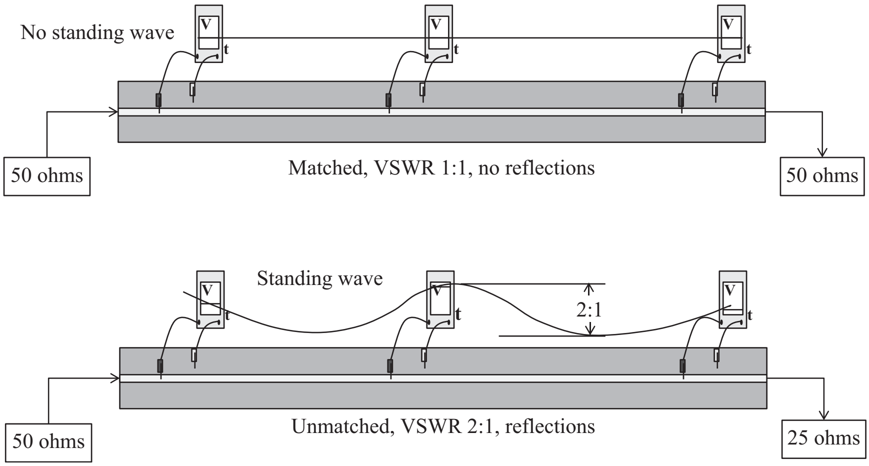


Figure 2.6 Unmatched creates a voltage standing wave

The first number defines the maximum voltage, and the second is the normalized minimum. The minimum is defined as one, and the maximum is the ratio of the two. The match between the PA and the antenna is often around 2:1, depending on the bandwidth of the antenna.

The standing wave is caused by reflections of the signal, due to a mismatch in the load, that are returned along the transmission line. The reflections are added by superposition, causing destructive and constructive interference with the incidence wave depending on the phase relationship between the incidence wave and the reflected wave. This results in voltages that are larger and smaller on points along the line. The main problem with mismatch and a high VSWR is that some of the power is lost when it is reflected to the source which is not delivered to the load or antenna. The standing wave is minimized by making the impedances equal so that there are virtually no reflections, which produces a VSWR of 1:1.

2.2.1 Maximum power transfer principle

The maximum power transfer principle states that for maximum power transfer to a load, the load impedance should be equal to the source impedance. For example, if the source impedance is equal to 50 Ω, then the load impedance should be equal to 50 Ω. This produces a VSWR of 1:1, which means that no power is reflected to the source; it is all delivered to the load (Figure 2.7).

However, the reverse is not true. If the impedance of the load is given, the source impedance is not chosen as equal to the load but is selected to minimal, or zero. Therefore, if the load impedance is given as 50 Ω, the source impedance is not selected as 50 Ω for maximum power transfer to the load but is chosen as 0 Ω. If the source is zero, then there is no loss in the source resistance (Figure 2.8). This applies especially for operational amplifiers (op amps) whose source impedance is low, around 10 Ω.

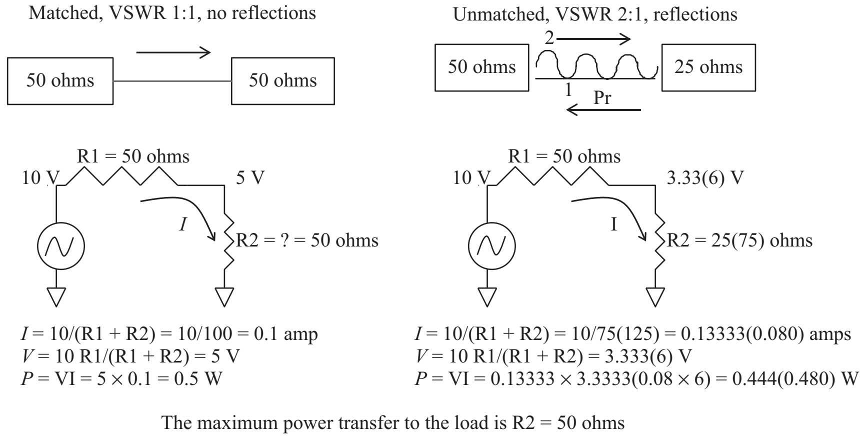


Figure 2.7 Maximum power transfer to the load with a given source impedance

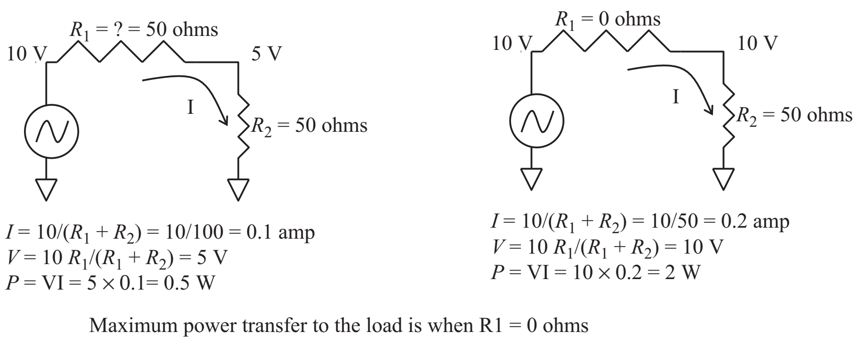


Figure 2.8 Maximum power transfer to the load with a given load impedance

2.3 Analog and digital communications

A modulator is used to transmit data information by attaching the data to a higher frequency carrier which is transmitted to the receiver using a small, practical antenna. For a typical analog system, the analog signal modulates the carrier directly by using either amplitude modulation (AM) or frequency modulation (FM) (Figure 2.9). The output of the modulator is a higher frequency signal that can be easily transmitted using a practical small antenna.

For a typical digital modulation system, the analog signal (e.g., voice) is digitized using a device known as an analog-to-digital converter (ADC). The ADC takes samples of the analog input and creates a digital pattern of "1"s and "0"s representing the value of the analog sample. This digital data stream is used to modulate the carrier frequency by shifting either the phase or frequency by the value of the digital signal (Figure 2.10). In addition, the modulator is used to generate spread spectrum and coding to enhance the digital signal.

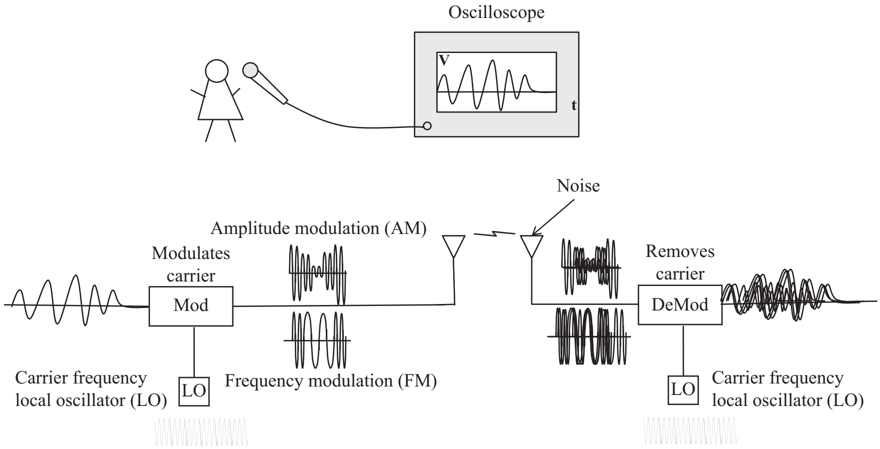


Figure 2.9 Baseband modulation for analog systems

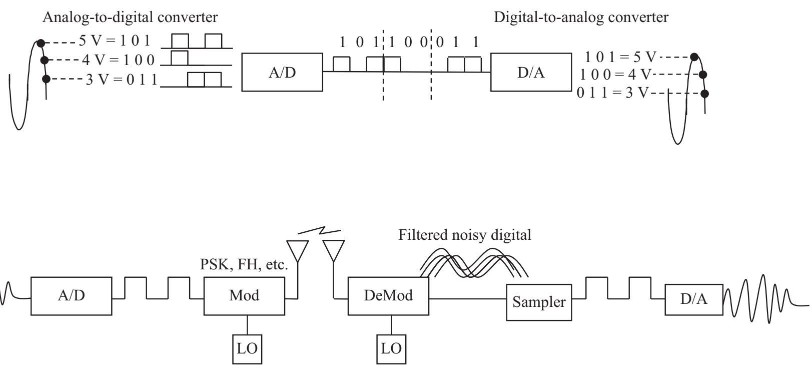


Figure 2.10 Baseband modulation for digital systems

Digital communications is becoming the standard for nearly all communications systems. The advantages of digital communications are as follows:

- Perfect reconstruction of the transmitted digital waveform assuming no bit errors. After digitizing the analog signal, the digital bit stream can be reconstructed at the receiver, and if there are no bit errors, it is an exact replica of what was sent by the transmitter. There is no gradual build-up as there is in analog systems, so the quality of the signal is excellent in a digital system until bit errors occur in the digital waveform (Figure 2.11).
- Smaller size due to custom application-specific integrated circuits and programmable chips such as digital signal processors (DSPs) and field-programmable gate arrays (FPGAs).

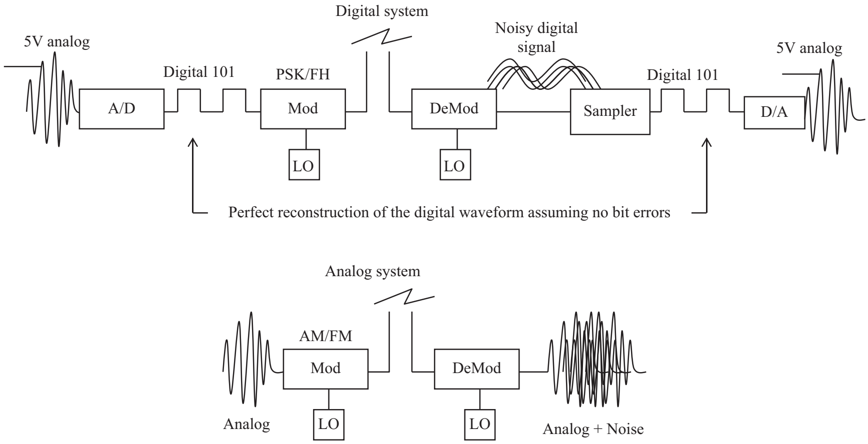


Figure 2.11 Digital versus analog communications

- Software-defined radios (SDRs) utilize the programmability of these DSPs and FPGAs to change several parameters including modulation to provide highly diverse radios that can be used for multiple applications and networking. Basically, it is like having multiple radios in one unit that is controlled by software.
- Cognitive radios (CRs) can be realized using digital communications and can be adaptive to the changing environment. The CRs can change functionality on the fly including frequency, modulation, bandwidths, and many other parameters made possible by the ability to program the digital circuitry to accommodate changes in the environment.

2.3.1 Digital-versus-analog communications

As mentioned before, digital transmissions can be perfectly reconstructed at the receiver, assuming there are no bit errors. However, when analog signals, such as voice, are used in a digital communication system, the analog signal needs to be digitized. This becomes a source of error, since this is only a digital approximation of the analog signal. With oversampling, this error becomes very small.

For an analog communication system, the actual analog signal modulates the carrier, and the analog signal is sent to the receiver. However, as noise and external signals distort the analog signal, it slowly degrades until it is undetectable or it reaches the minimum detectable signal (Figure 2.11).

The advantage of a digital signal is that as it degrades, the quality of the signal is not affected, assuming the receiver can still detect the bits that were sent from the transmitter with no bit errors. Therefore, it does not matter how noisy the signal is coming into the receiver, if the bits can be detected without bit errors or the errors can be corrected, then the quality of the signal does not change or sound noisy.

An example of this is the comparison between analog cell phones and digital cell phones. Analog cell phones continue to get noisier until it is difficult to discern

the voice because of so much noise, whereas digital cell phones are perfectly clear until they start getting bit errors, which results in dropouts of the voice or missing parts of the conversation.

2.3.2 Software programmable radios and cognitive radios

Software programmable radios are often referred to as SDRs. A big advantage of digital communications is that the modulation and demodulation can be accomplished using software. This makes the design much more flexible since this technique is used to change to different modulation schemes. The principle is that the software is programed in both the transmitter and the receiver to provide multiple modulation schemes that can be selected on a real-time basis. If there are multiple transmitters with different modulation types, the receiver can be programed to accept these different modulation types by changing the software. An example of this is using a data link that is designed and built to transmit and receive a waveform called AM minimum shift keying (AM-MSK). By simply changing the software or changing the software load of either a DSP or FPGA or both, this modulation can change to differential 8-phase-shift keying (D8PSK). This was accomplished in setting the standards for special category 1 landing systems. The software can be changed at the factory, or for a real-time system, the software loads can reside in memory on the data link and the desired software can be loaded on command of the data link.

The hardware uses DSPs, FPGAs, or both. FPGAs are generally used for high-speed processing, while DSPs are generally used for low-speed processing or general processing tasks. Many data link systems today use both FPGAs and DSPs and partition the functionality between these parts. In addition, hardware manufacturers are designing product that incorporates both functions. This technique can be used to optimize the function of the data link. A basic SDR is shown in Figure 2.12.

CRs are an extension of SDRs. They use smart radios that can sense the present environment for existing radios with given modulations and data links and program the transmitter, receiver, or both for interoperability. In addition, CRs can automatically determine which bands or parts of bands have the minimum amount of noise, jammers, or other users and incorporate this information to determine the best band for operation of the cognitive data link. This maximizes the number of users by optimizing the data link with respect to frequency, modulation, time, and space and provides the best noise-free and jammer-free operation.

2.4 Digital modulation

The most common form of digital modulation is changing the phase of the carrier frequency in accordance with the digital signal that is being sent. This is known as phase-shift keying (PSK) or direct sequence digital modulation. This digital sequence can be either the digitized data or a combination of digitized data and a spread spectrum sequence. There are many different levels and types of PSK; only

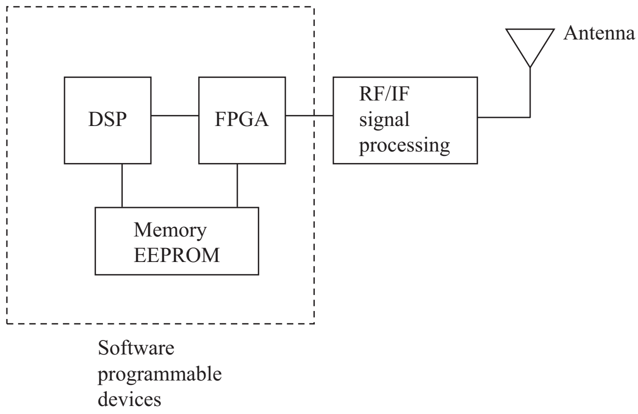


Figure 2.12 Basic software defined radio

basic modulation methods are discussed here. However, the principles can be extended to higher order PSK modulations. The basic form of digital communication is shown in Figure 2.13. This shows the carrier frequency and the digital data being fed into a modulator. The modulator is binary PSK (BPSK) and changes the phase between 0° and 180° according to the digital data being sent.

2.4.1 Binary phase-shift keying

The BPSK modulator and phasor diagram are shown in Figure 2.14. BPSK is defined as shifting the carrier 0° or 180° in phase, depending on the digital waveform. For example, a binary "0" gives 0° phase of the carrier, and a binary "1" shifts the carrier by 180°. To produce the digital waveform, the data or information signal is digitized, encoded, and sent out in a serial bit stream (if not already). The end result is a serial modulating digital waveform representing the data to be transmitted. This digital output contains 0 and 1 and often needs to be changed to ± 1 to directly modulate the carrier frequency in a typical mixer application. However, certain forms of hardware can bypass this step and modulate the mixer directly or phase shifter. The output of the mixer or phase shifter is a BPSK modulated carrier signal that is transmitted and sent over to the receiver for demodulation and detection. As the carrier phase changes 180°, the hardware does not allow for an instantaneous change in phase, so the amplitude goes to zero, which produces 100% AM (Figure 2.14). This is called the zero crossover point and is an unwanted characteristic of BPSK and can cause degradation to the waveform and performance.

A high-speed pseudonoise (PN) code and modulo-2 adder/exclusive-or functions are added to the basic BPSK modulator to produce a spread spectrum waveform (Figure 2.15). The high-speed code generates a wider spectrum than the data spectrum needed for communications, which is known as a spread spectrum.

Applying minus voltage to an actual RF mixer reverses the current through the balun of the mixer and causes the current to flow in the opposite direction, creating

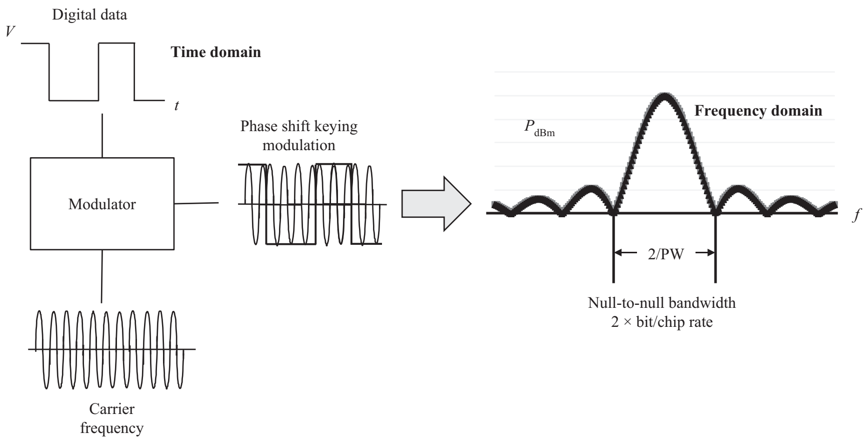


Figure 2.13 Shifts the phase of the carrier frequency according to the data

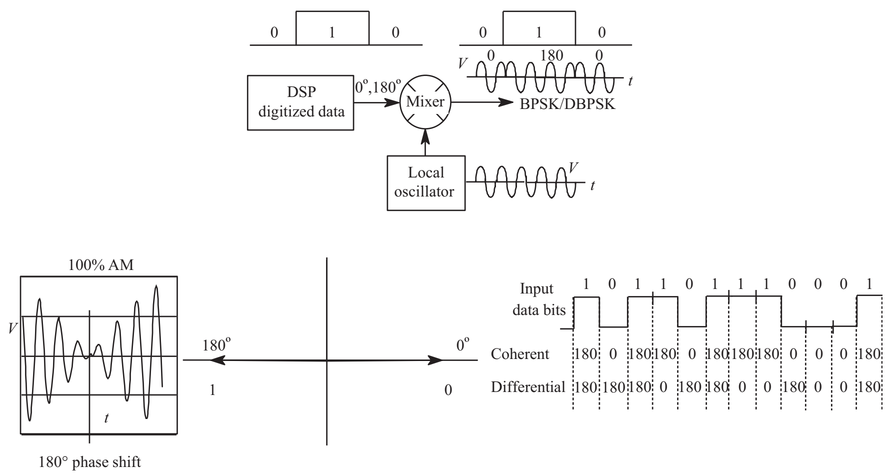


Figure 2.14 BPSK generator

a net 180° phase shift of the carrier. Thus, the carrier is phase shifted between 0° and 180° depending on the input waveform. A simple way of generating BPSK spread spectrum is shown in Figure 2.14. The LO is modulated by this digital sequence producing a 0° or 180° phase shift on the output of the mixer. Other devices such as phase modulators and phase shifters can create the same waveform as long as one digital level compared with the other digital level creates a 180° phase difference in the carrier output. Other digital methods incorporate the phase shifter in the digital hardware and firmware.

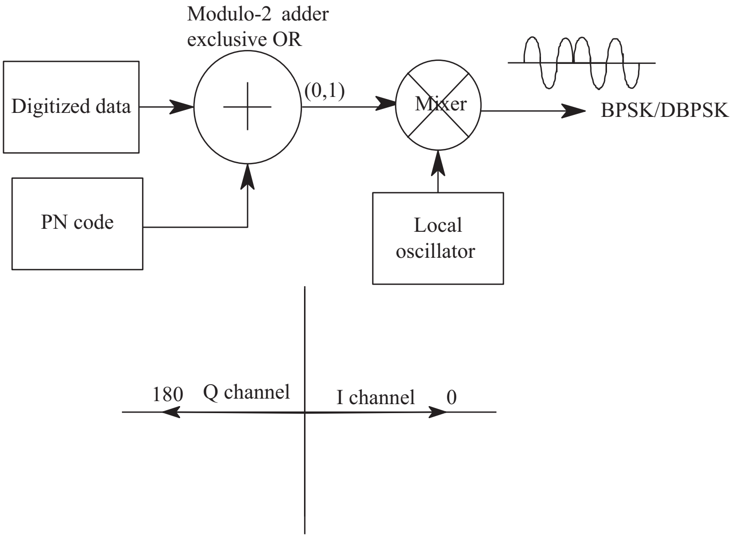


Figure 2.15 BPSK spread spectrum generator

2.4.2 Quadrature phase-shift keying

Quadrature PSK (QPSK) is generated by quadrature phase shifting the LO so that four possible phase states are produced at the output. One method of producing these phasors is by using two channels: one channel containing an LO that is in phase, and the output of the mixer is BPSK modulated to produce 0° or 180°; and the other channel containing the same LO that is shifted by 90° so that the output of the mixer is BPSK modulated to produce 90° or 270°. These two channels are then combined to produce the four phase states (Figure 2.16). The two BPSK systems are summed together, which gives four possible resultant phasors −45°, 135°, −135°, and −45°, which are all in quadrature. Since the digital transitions occur at the same time, changes between any four resultant phasors can occur.

Depending on the input of both bit streams, the phase of the resultant can be at any of the four possible phases. For example, if both bit streams are 0, then the phasor would be 45°. If both bit streams changed to 1, then the phasor would be at −135°, giving a change of carrier phase of 180°. If only the first channel changes to 1, then the phasor would be at −45°, giving a change of 90°. Therefore, the four possible phase states are 45°, 135°, −135°, and −45°. The phasor diagram can be rotated, since it is continually rotating in time and only a snapshot is shown, and then the phasors would be at 0°, 90°, −90°, and 180° for QPSK generation. QPSK has the capability of 180° phase shifts depending on the code and the previous phase state, so it often goes through the zero crossover point which produces 100% AM during the phase transition as shown in Figure 2.16. This causes unwanted characteristic of QPSK and can cause degradation to the waveform and performance.

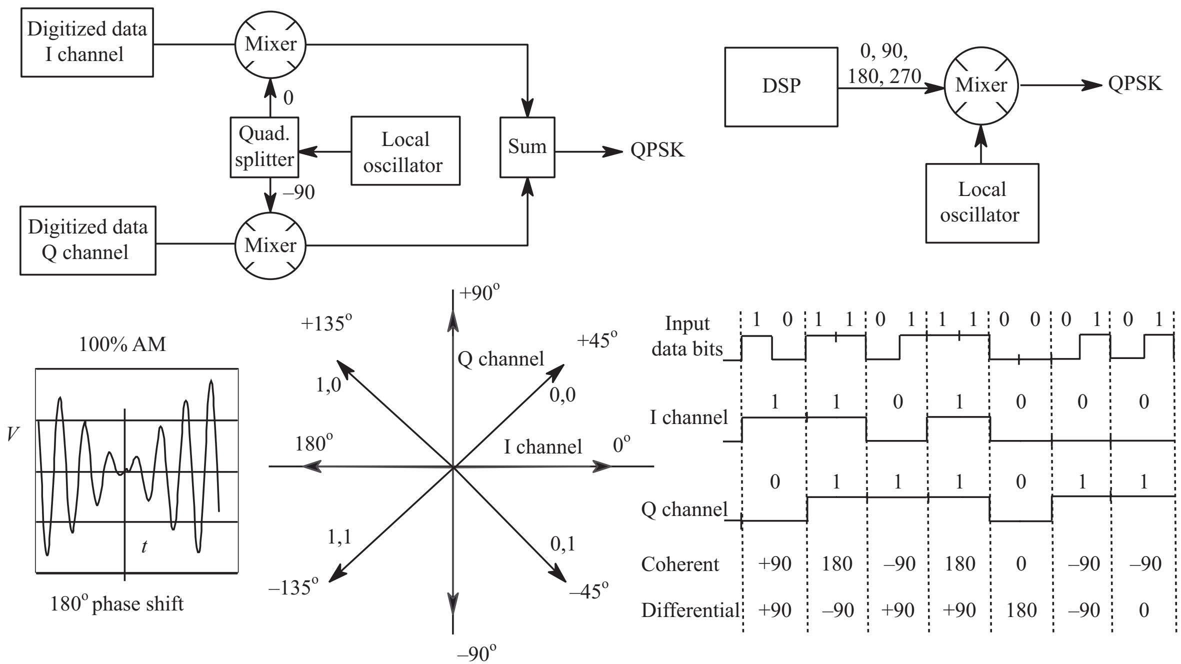


Figure 2.16 QPSK generator

Usually, the LO contains the 90° phase shift that is used to provide the two quadrature channels of the BPSK phasors instead of trying to provide the 90° phase shift of the actual binary input. Either way would provide the 90° quadrature phase rotation that is required. However, quadrature rotation of the binary input requires phase shifting all of the frequency components by 90°. This requires a more sophisticated and expensive phase shifter that is broadband. Shifting only the carrier to produce the quadrature channels requires a phase shift at one frequency, which is much simpler to build. In fact, in the latter case, a longer piece of cable cut to the right length can provide this 90° quadrature phase shift. As mentioned before, these phase shifting operations can be accomplished directly in the digital and software functions.

2.4.3 Offset QPSK

Another type of QPSK is referred to as offset QPSK (OQPSK) or staggered QPSK. This configuration is identical to QPSK except that one of the digital sequences is delayed by a half-cycle so that the phase shift occurs on only one mixer at a time. Consequently, the summation of the phasors can result in a maximum phase shift of only ±90° and can change only to adjacent phasors (no 180° phase shift is possible). The phasor diagram in Figure 2.17 is identical to Figure 2.16, except no 180° phase shift is possible. This prevents the zero-amplitude crossover point and provides smoother transitions with less chance of error, reducing the AM effects common to both BPSK and QPSK (Figure 2.17). OQPSK changes only 90° so that the phasor AM is much less (approximately 29%). Figure 2.17 also shows a simple way of generating OQPSK. The only difference between this and the QPSK generator is the 1/2-bit delay in one of the bit streams. This prevents the 180° phase shift present in the QPSK waveform.

2.4.4 Higher order PSK

The previous analysis can be extended for higher order PSK systems. The same principles apply, but they are extended for additional phase states and phase shifts.

Direct sequence waveforms are used extensively for digital modulation systems. BPSK is the simplest form of PSK, shifting the carrier 0° or 180°. QPSK uses two BPSK systems, one in quadrature, to achieve four phase shifts: 0°, 90°, −90°, and 180°. OQPSK is used to minimize AM and is identical to QPSK with the phase transitions occurring at 1/2-bit intervals and at different times for each of the BPSK channels. This provides phase shifts of 0°, 90°, and −90° degrees and eliminates the 180° phase shift. Higher order PSK systems can be analyzed much the same way, only with more phase states and phase transitions. However, with more phase shift possibilities it becomes harder to detect and resolve the different phase states. Therefore, there is a practical limit on how many phase states can be sent for good detection. This limit seems to grow with better detection technology, but caution must be observed in how many phase states can be sent out reliably for practical implementation.

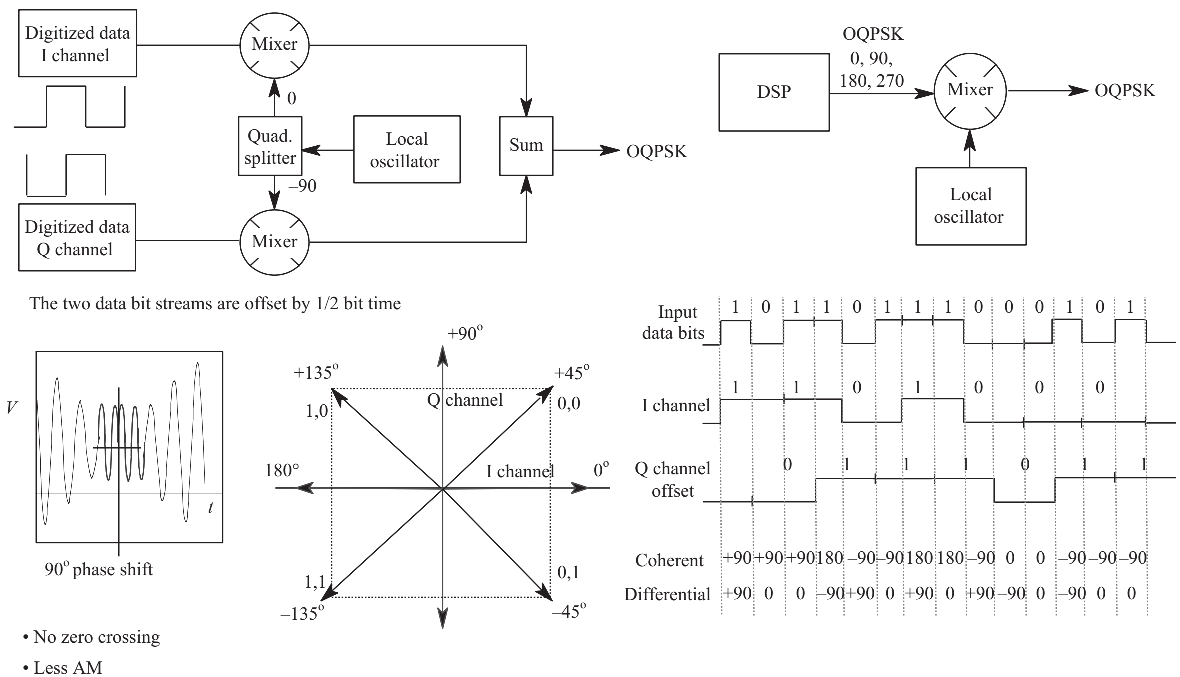


Figure 2.17 OQPSK generator

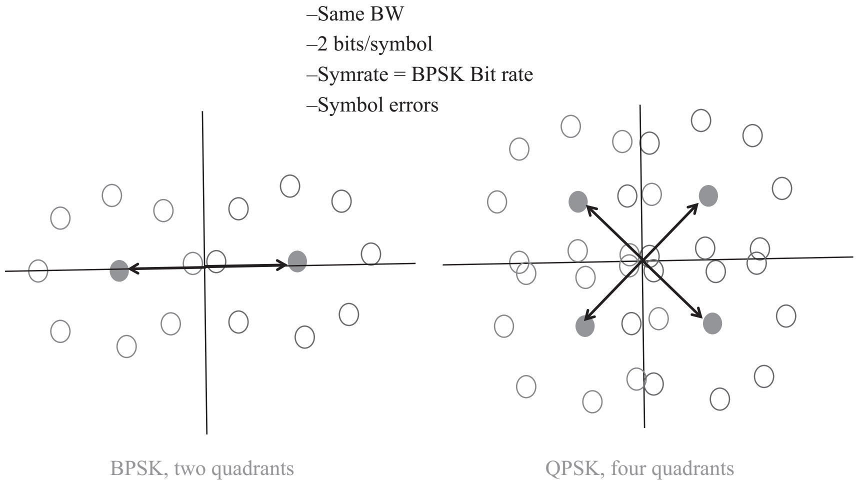


Figure 2.18 Constellations of BPSK and QPSK with the same signal power

2.4.5 BPSK versus QPSK constellation comparison

BPSK uses two quadrants or two phase states to transmit the data. QPSK, on the other hand, uses four quadrants or four phase states, each state representing 2 bits of data, referred to as symbol rate. Therefore, QPSK can send twice as many data bits as BPSK. However, using four quadrants increases the bit error rate since the symbols do not have enough separation as shown (Figure 2.18).

This can be overcome by either increasing the signal power or reducing the noise by decreasing the bandwidth as shown (Figure 2.19). Reducing the bandwidth requires reducing the symbol rate by 1/2, so the bit rate for both BPSK and QPSK is the same

$$\text{BPSK: Pe} = 1/2\text{erfc}\left((E_b/N_o)^{1/2}\right)$$
$$\text{QPSK: Pe} = 1/2\text{erfc}\left((E_b/N_o)^{1/2}\right) - 1/4\text{erfc2}\left((E_b/N_o)^{1/2}\right)^*$$

* The second term can be eliminated if $E_b/N_o \gg> 1$.

Increasing the signal power allows QPSK to send twice the data as BPSK in the same bandwidth.

2.4.6 8-Level PSK

The 8-PSK type of modulation includes the phase possibilities of 0°, 90°, −90°, and 180°, thus providing 0°, 45°, −45°, 90°, −90°, 135°, −135°, and 180° phases. This provides eight possible phase, or 3 bits (2^3) of information (Figure 2.20).

Thus, for the same bandwidth as BPSK, 8-PSK can send three times as many bits (i.e., the bit rate is three times the bit rate of BPSK); however, 8-PSK requires a higher E_b/N_o. The actual phase shifting occurs at the same rate as BPSK. The rate

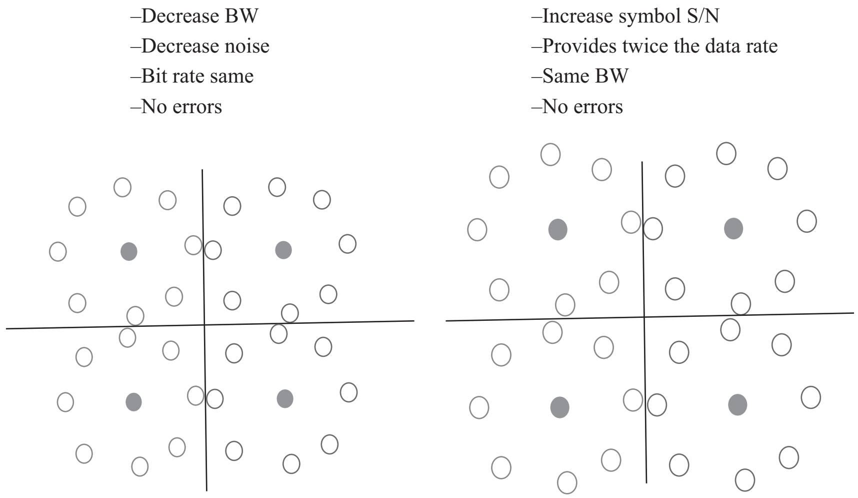


Figure 2.19 Improvements by decreasing bandwidth or increasing S/N for QPSK

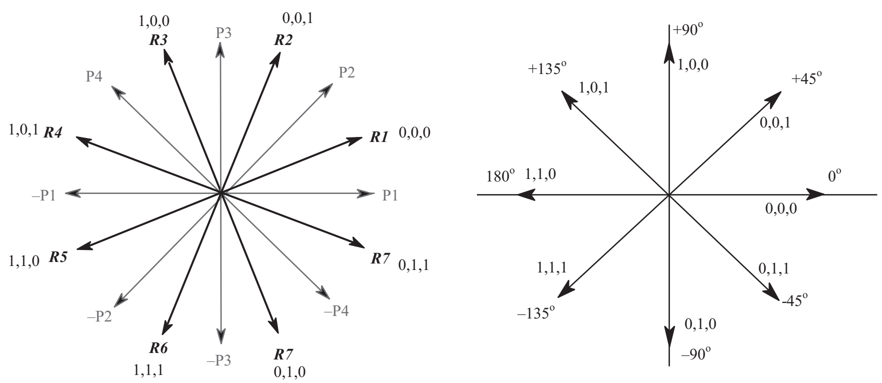


Figure 2.20 Phasor diagrams for 8 phase shift keying (8PSK) modulation

of the phase shifts is called the symbol rate. The symbol is the rate at which the phase is shifted, but the actual bit rate in this case is three times the symbol rate since there are 3 bits of information in each phase shift or symbol. The symbol rate is important because it describes the spectral waveform of the signal in space. For example, if the symbol rate is 3 ksps (3,000 samples per second), then the null-to-null bandwidth would be 6 kHz wide (two times the symbol rate), which would decode into 9 kbps (data rate). Note that this type of modulation allows a 180° phase shift, which experiences AM.

2.4.7 π/4 differential QPSK

The π/4 differential QPSK (DQPSK) is a modulating scheme that encodes 2 bits of data as ±45° and ±135° phase shifts that are in quadrature. The next phase shift will be from that resultant phasor which makes it differential since it depends on the previous level and looks from a difference in phase. Therefore, if the first phase shift is +135°, and the second phase shift is +135°, the resultant phasor's absolute phase would be at 270°. Because it is differential, once the phase shift occurs, the next phase shift starts at the last phase state. Then, according to the data bits, it shifts the phase either ±45° or ±135°. Another example: if a +45° shift occurs referenced to the last phasor sent, then the bits that are sent equal 0,0. Proceeding with the other phase shifts, a −45° shift is 0,1; a +135° shift is 1,0; and a −135° shift is 1,1 (Figure 2.21). The shifts for the next bits sent start from the last phasor measured. Over time, this will eventually fill up the constellation and will look similar to the 8-PSK diagram. Note that none of the phase shifts go through zero amplitude, so there is less AM for this type of modulation compared with QPSK, which does go through zero amplitude.

2.4.8 16-Offset quadrature amplitude modulation

Another common modulation scheme is 16-offset quadrature AM (16-OQAM), which is very similar to OQPSK. In 16-OQAM, each of the phasors has two amplitude states before summation. The resultant phasor in one quadrant has four possible states (R1, R2, R3, and R4), as shown in the phasor diagram (Figure 2.22).

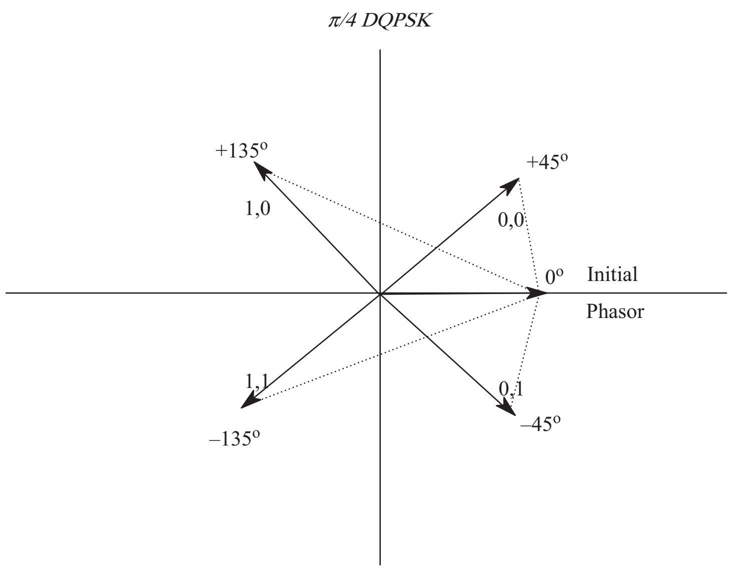
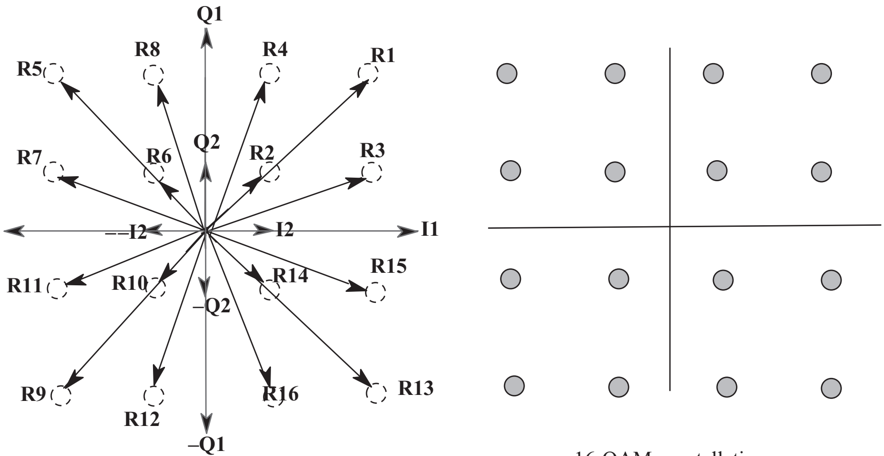


Figure 2.21 Phasor diagrams for a π/4 differential QPSK modulation



16-QAM phasor diagram

16-QAM constellation
displayed using a vector analyzer
max. separation for noise

Figure 2.22 Amplitude/phasor diagram for 16 OQAM

Offset is used here for the same reasons as in OQPSK—to prevent transitions through zero amplitude and to reduce the AM on the output.

Since there are four possible amplitude/phase positions in one quadrant and a total of four quadrants, there are sixteen possible amplitude/phase positions in this modulation scheme (Figure 2.22). These 16 states equates to 4 bits of information for each change in state or symbol, which provides four times the data rate compared with BPSK. However, a higher E_b/N_o is required.

Since this is 16-OQAM, the "offset" means that the phasors in the defined quadrant can change only to the adjacent quadrants. For example, the phasors shown in quadrant 1 can change only to quadrants 2 and 4. These phasors cannot change from quadrant 1 to quadrant 3. In other words, only the I or Q channel can change in phase at one time, not both at the same time (Figure 2.22).

2.4.9 Phasor constellations and noise immunity

A phasor constellation is a technique used to determine the quality of phase modulated waveforms and to distinguish between phase states. A phasor constellation shows the magnitude and phase of the resultant phasor for BPSK and 16-QAM (Figure 2.23). BPSK has two states which provides 1 bit of data. 16-QAM has 16 states which provide 4 bits of data. A vector analyzer is used to display the constellations where the phasors appear as points in a two-dimensional display representing both phase and amplitude and with no noise in the system. These constellation points show the amount of separation between one phasor state point and another.

To understand the noise immunity of different modulation types, Figure 2.24 shows two types of phase modulations, including noise, which varies the phasor's

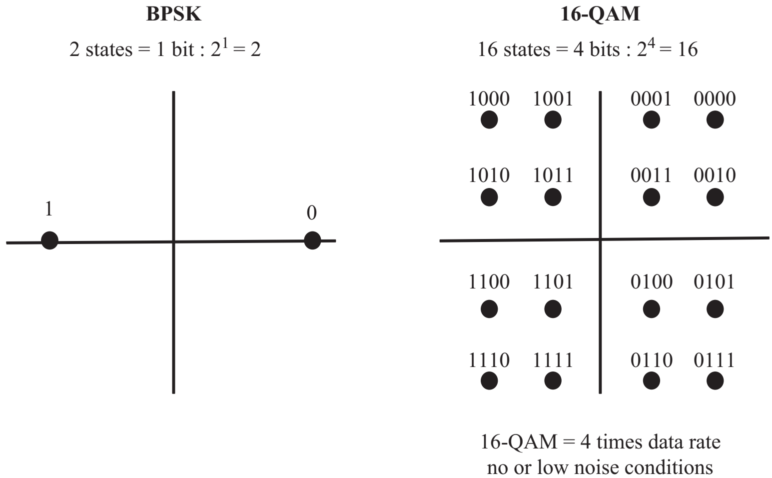


Figure 2.23 Higher order modulation results in higher data rates

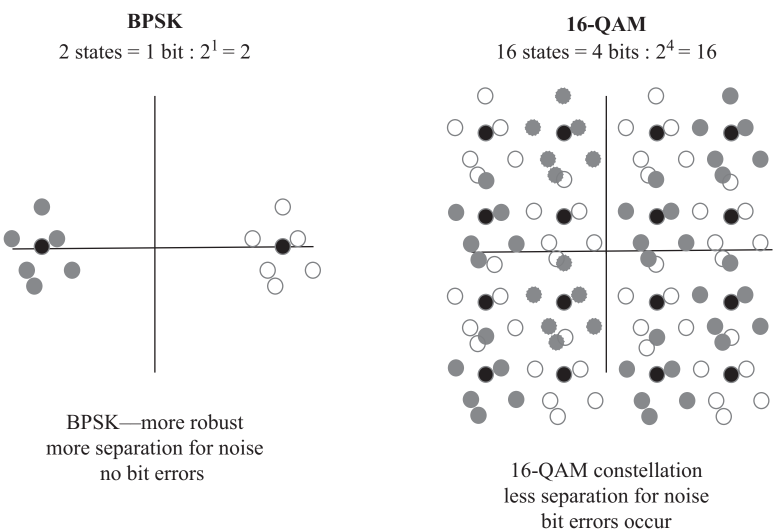


Figure 2.24 Constellation and noise immunity between BPSK and 16-QAM

end points for BPSK and 16-OQAM modulated signals. For BPSK, the constellations are close enough together that it is very easy to detect which value of phase was sent, 0° or 180°. Therefore, BPSK is much more resistant to noise and variations in phasor values. Although the 16-OQAM signal contains more bits per phasor value or symbol, which provides a faster data rate for a given bandwidth, the actual phasor value is more susceptible to noise, and it is much more difficult to

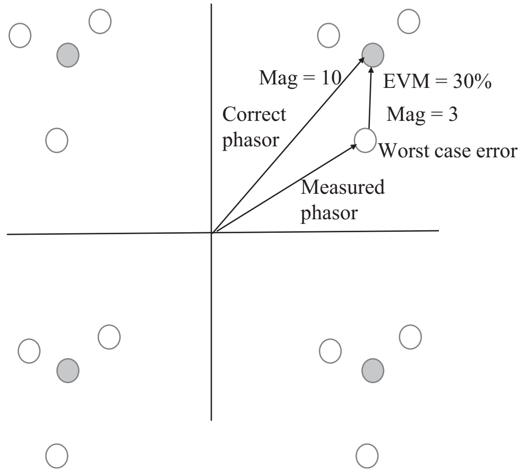


Figure 2.25 Error vector magnitude (EVM)

separate the noisy phasor values, which results in more bit errors. This is a trade-off between noise resistance and higher data rates.

The quality of the phasor values can also be measured by determining the error vector magnitude (EVM). An error vector is a vector in the I–Q plane between the ideal vector or constellation point and the received vector or constellation point (Figure 2.25). The constellation points are the ends of the vectors and appear as points on a vector analyzer. These points represent the magnitude and phase of the vectors. The average length or magnitude of the error vector, defined as the distance between these two points, is the EVM. The ratio of the EVM (the distance between constellation points) and the ideal vector magnitude multiplied by 100 provides the EVM as a percentage. For example, if the ideal phasor magnitude is equal to 10 and the EVM is equal to 3, then the EVM in percent would be $3/10 \times 100 = 30\%$ EVM (Figure 2.25). The smaller the percentage EVM, the less is the noise and the higher the probability of detecting the correct signal phase and amplitude.

2.4.10 Continuous phase PSK

Continuous phase PSK or CP-PSK is a sinusoidal weighted O-QPSK waveform that changes phase following sinusoidal transitions around the unit circle which keeps the magnitude constant during the phase shift. This does not allow zero amplitude crossovers between phase shifts which prevents any AM and is therefore immune to saturation effects. This also provides improved spectral efficiency and prevents spectral regrowth or sidelobe regeneration due to nonlinearities in the hardware. CP-PSK modulation is being used extensively in data links due to spectral characteristics, burst-type modulation, and packet radios, where data are sent in bursts or packets. The modulation causes the phase of the carrier to shift to

another state, where it can remain in that given state for a period of time (Figure 2.26).

CP-PSK can be created by sinusoidal weighting quadrature channels, I and Q channels, and then combining the channels for the modulation output (Figure 2.27). This shows the envelopes of the I and Q channels, and the final constant envelop (no AM) continuous PSK.

2.4.10.1 Spectral efficiency

CP-PSK has much better spectral efficiency with reduced sidelobes to prevent unwanted out-of-band power than other types of modulations such as BPSK and QPSK (Figure 2.28). This provides more power in the main lobe for detection and reduces interference to the out-of-band signals and users. In addition, this modulation is more low probability of intercept/low probability of detect (LPI/LPD) since there is lower power spread across the spectrum.

2.4.10.2 Spectral regrowth

Digital modulation causes sidelobes to be generated by the digital signal. These sidelobes extend continuously in the frequency domain. In order to prevent interference with other out-of-band signals, to comply with FCC regulations, and to reduce probability of detection, these sidelobes need to be filtered. However, nonlinear functions after filtering can cause spectral regrowth or sidelobe regeneration. The nonlinear devices include PAs such as class C amplifiers, limiters, and any devices that are operated in compression/saturation. This is basically a component or components that are driven into saturation after filtering. Modulation waveforms that exhibit AM which includes BPSK, QPSK, and others are prone to spectral regrowth (Figure 2.29). For example, after the sidelobes are filtered, when they are amplified by the PA—which most of the time is nonlinear—the sidelobes reappear. This can be a major concern when complying with regulations and preventing out-of-band interference. Continuous phase modulation schemes such as CP-PSK minimize this spectral regrowth problem and keep the sidelobes down even when they are exposed to nonlinear devices. Since CP-PSK is a constant envelope with no AM, it eliminates spectral regrowth or sidelobe regeneration (Figure 2.29).

2.4.10.3 Summary of advantages using CP-PSK

1. Spectrally efficient, reduced sidelobes, and out-of-band energy.
   (i) BPSK/QPSK—13 dBc for first sidelobe, 6 dB per octave rolloff.
   (ii) CP-PSK—23 dBc for first sidelobe, 12 dB per octave rolloff.
2. Operates in saturation—constant amplitude, no AM
   (i) Ideal for radar bursts that operate in saturation.
   (ii) Ideal for small power limiting applications such as missiles, UAVs for low power, small size, low-cost class-C PAs operated in saturation.
3. Reduces intersymbol interference (ISI), minimizes bit errors.
4. No sidelobe regeneration or spectral regrowth.

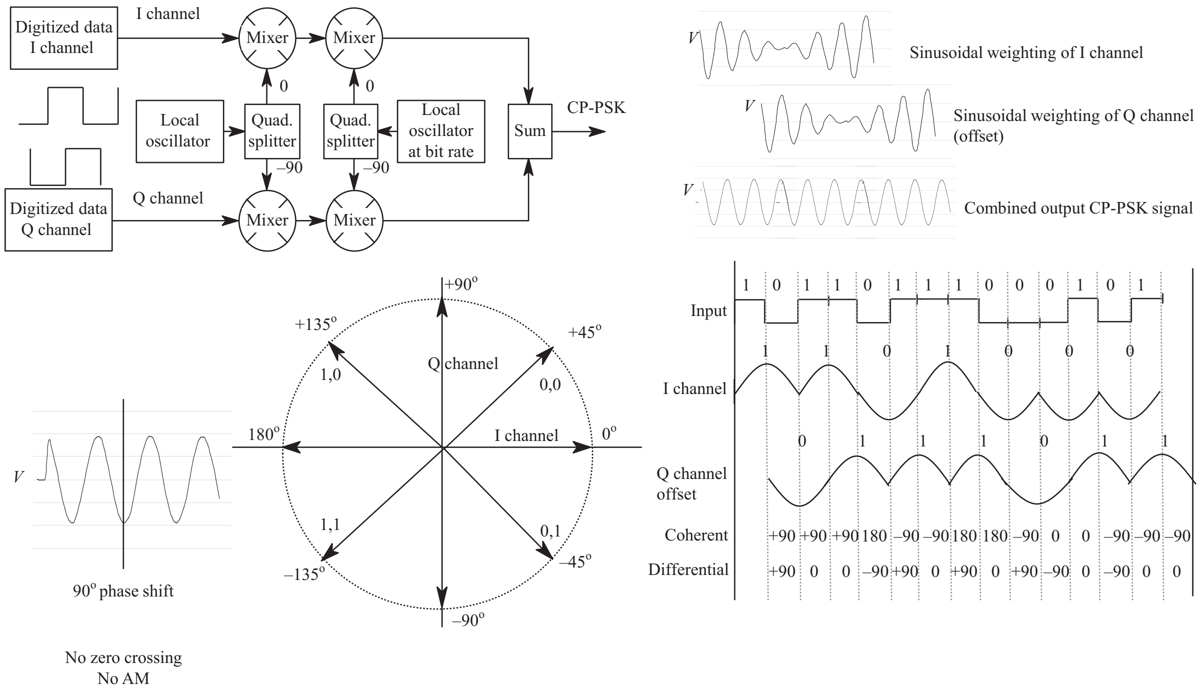


Figure 2.26 Continuous phase-phase shift keying (CP-PSK)

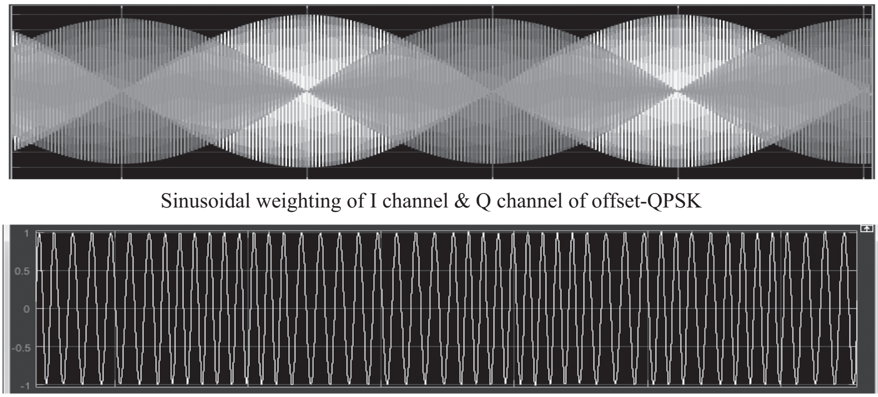


Figure 2.27 Continuous phase-phase shift keying (CP-PSK) modulation

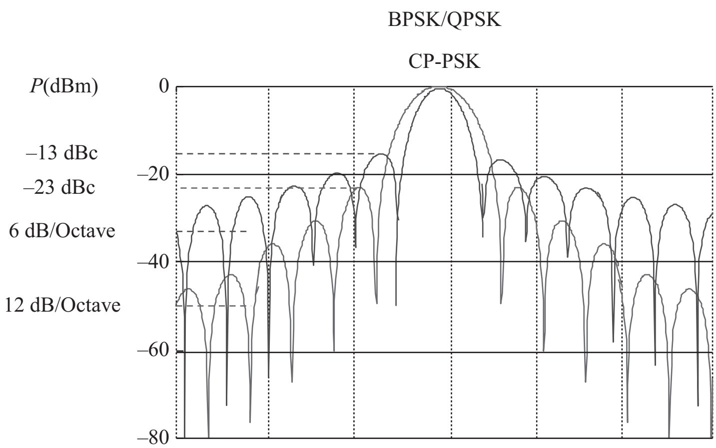


Figure 2.28 Out-of-band sidelobes are suppressed with CP-PSK

 (i) BPSK/QPSK filtered sidelobes reappear after going through any non-linear device such as PAs.
 (ii) CP-PSK filtered sidelobes do not regenerate.

5. Provides coherency for burst communications.
 (i) Maintains the phase during off times.
 (ii) Longer dwells on phase state for more robust detection.

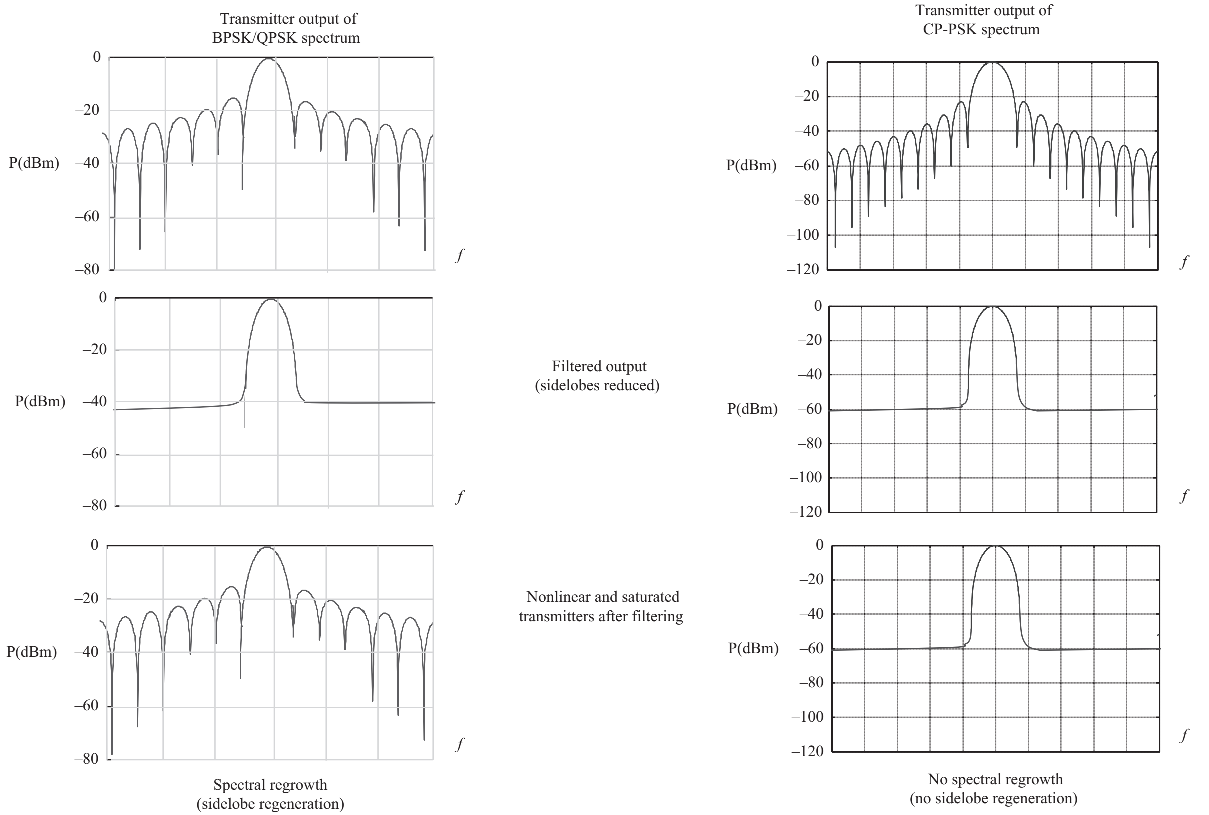


Figure 2.29 CP-PSK prevents spectral regrowth or side lobe regeneration

2.4.11 Summary of PSK modulations

Many other PSK configurations are in use today, and there will be other types of modulating schemes in the future. One of the main concerns with using PSK modulators is the AM that is inherent when phase states are changed. BPSK switches from 0° to 180°. Since this is not instantaneous, due to bandwidth constraints, the phasor passes through zero amplitude in the process. The more band-limiting in the receiver, the more AM is present in the resultant waveform. The magnitude during the phase change goes to zero, thus creating a 100% AM of the carrier. This causes problems in saturation, increases ISI, and is susceptible to sidelobe regeneration. In saturation, the zero amplitude (noise only) is increased to the saturation level which causes problems in detection of the phase transitions.

QPSK also has the same problem when the phase switches from 0° to 180° (Figure 2.30); however, it does have a slight advantage over BPSK since during the 90° phase changes, the AM is reduced.

Various schemes have been developed to reduce this AM problem. For example, OQPSK or O-QPSK prevents the 180° phase change which allows only a maximum phase shift of 90° so that the AM problem is significantly reduced. However, there is still AM which can still cause problems in saturation, but not as pronounced as the BPSK and QPSK modulations (Figure 2.30).

Moreover, MSK is used to smooth out the phase transitions, which further reduces the AM problem and makes the transition continuous. CP-PSK, also referred to as constant envelope modulation, is a sinusoidal weighted O-QPSK which changes phase following the unit circle. This keeps the magnitude constant during the phase shift and prevents AM. The phase shifts are continuous around the phasor diagram with no change in amplitude and is therefore immune to saturation effects (Figure 2.30). It is also generated by minimizing frequency shift keying frequency-shift keying (FSK).

2.4.12 Differential phase-shift keying

The BPSK waveform above can be sent out as absolute phase (i.e., a 0° phase shift is a "1," and a 180° phase shift is a "0"). This type of system is known as a coherent system, or coherent BPSK. Another way to perform this function is to use differential PSK (DPSK), which generates and detects a change of phase. A change of phase (0° to 180° or 180° to 0°) represents a "1," and no change (0° to 0° or 180° to 180°) represents a "0."

Differential techniques can be applied to higher order PSK systems such as DQPSK and D8PSK. It is also used to create continuous phase-PSK (CP-PSK).

2.4.12.1 Advantages and disadvantages of differential modulation

Coherent modulation maintains a constant state throughout the signal. To achieve this, it requires a very stable source and phase locking techniques. It is generally more costly and more complex. On the other hand, differential modulation is less complex and less expensive. It needs to be stable only between state changes

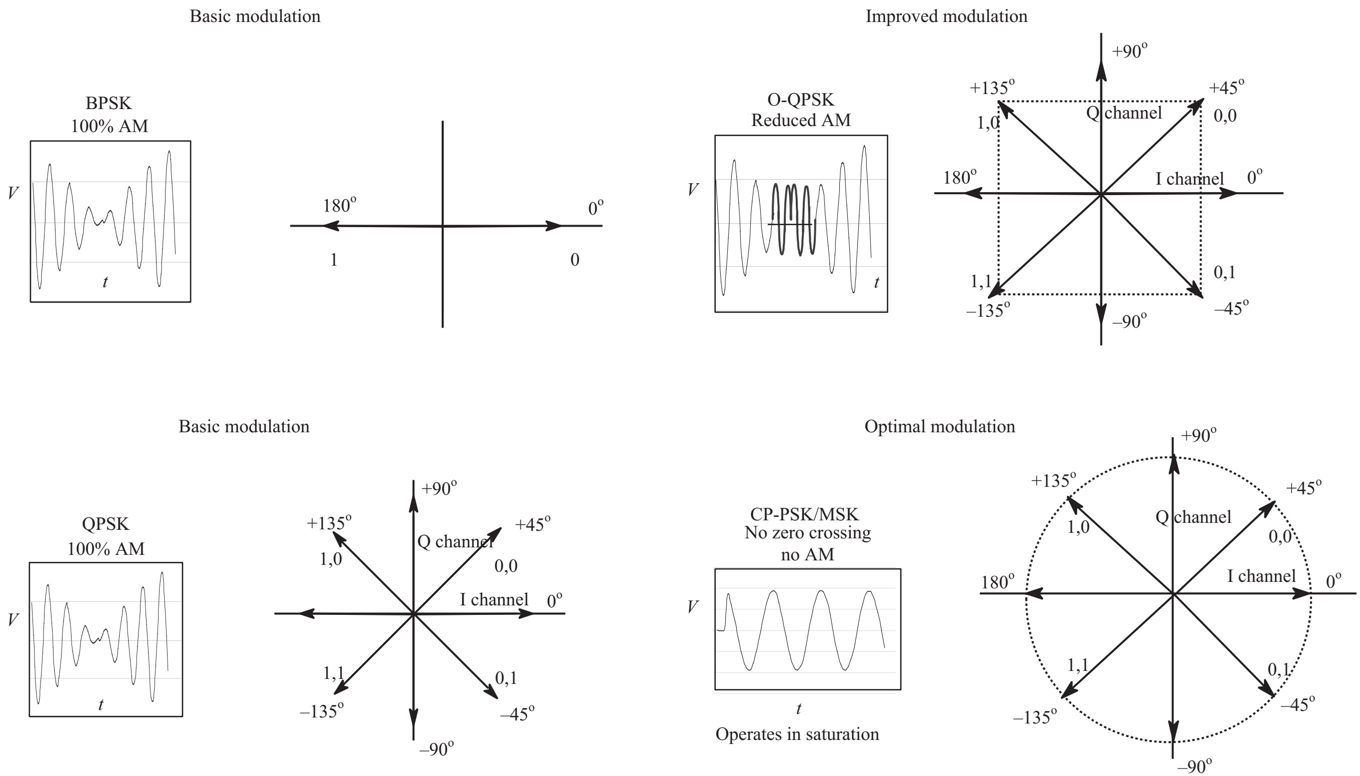


Figure 2.30 Phasor and time diagrams for different types of modulations

(phase/amplitude) of adjacent symbols. Thus, stability is critical just for a short period of time. However, differential modulation creates more bit errors and requires more signal-to-noise ratio (SNR) compared with coherent systems.

A diagram showing the differences between coherent and differential systems with respect to the number of bit errors is shown in Figure 2.31. For coherent systems, if there is a bit error, it affects only the bit that was in error. For differential systems, there is the possibility of having twice as many errors. If the system receives 1 bit error, then the next bit could be in error, since it is dependent on the previous bit value and monitors the amount of change to the bit (Figure 2.31).

There are several advantages using differential modulation and is generally the best type of modulation for most systems. These advantages generally outweigh the small increase in bit errors.

One major advantage of the differential system is the simplicity and cost of implementing a demodulation scheme. For example, a simple differential BPSK system can be demodulated by using a delay and multiply, as shown in Figure 2.32. The delay is equal to 1 bit time, and the multiplication and integration produces the bit that was sent. When the signal is in phase, then the delay and multiply detection integrates to a large positive number; when the signal is out of phase, the signal integrates to a large negative number, thus retrieving the differentially encoded bit stream (Figure 2.32).

Another big advantage for using differential modulation is it is much easier to detect because the absolute phase does not need to be determined; just the change of phase is monitored. This makes this technique less costly due to the demands of keeping the phase constant over a longer period of time. There are many factors that can change the stability of the phase measurement including the effects of Doppler, oscillator instability, phase noise, and other phase disturbances. However, DPSK makes the phase decision between adjacent phase shifts and is only depending on the previous phase to decide phase change. For fast moving platforms including satellite communications, the effects of Doppler are minimal.

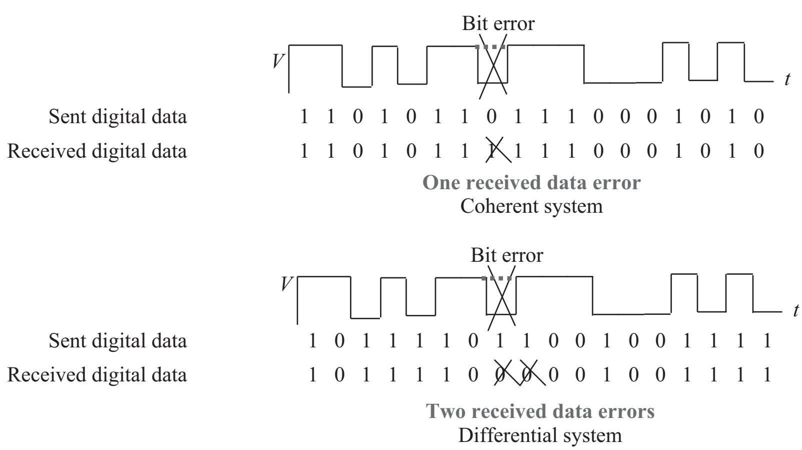


Figure 2.31 Coherent versus differential systems with respect to bit errors

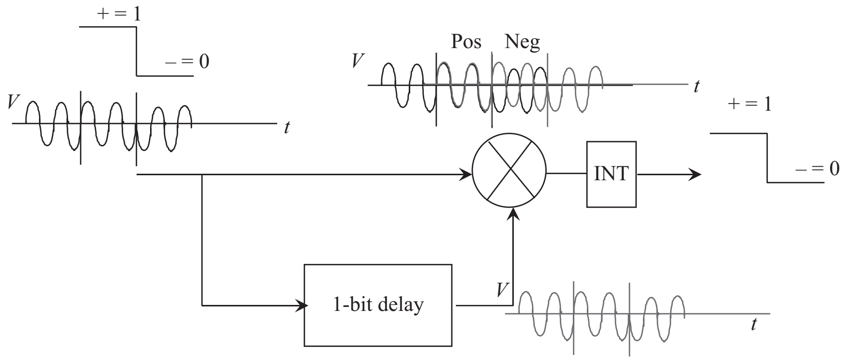
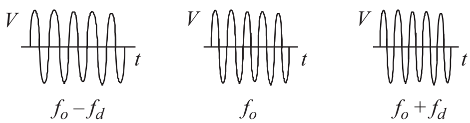


Figure 2.32 Differential demodulation using delay and multiply

The following analysis shows that Doppler can be reduced using differential modulation:

Doppler Frequencies: $f_t = f_o - f_d$ or $f_o + f_d$: depends on away or towards worst case



Example 1: Given: $v = 1{,}000$ m/s, $f_o = 10$ GHz, $c_o = 3 \times 108$ m/s, BR = 150 Mbps
$f_d = v \times f_o/c_o = (1{,}000$ m/s $\times$ 10 GHz)/3 $\times$ 108 m/s = 33 kHz
Differential Bit time $t_d = 1/150$ Mbps = 6.67 ns
Degree change (error) due to Doppler over t_d
$\Delta° = 33$ kcyc/s $\times$ 360°/cyc $\times$ 6.67 ns = 0.08° error
%error = 0.08°/180° $\times$ 100 = 0.04% for D-BPSK
%error = 0.08°/90° $\times$ 100 = 0.09% for D-QPSK, CP-PSK

Example 2: Given: $n = 1{,}000$ m/s, $f_o = 30$ GHz, $c_o = 3 \times 108$ m/s, BR = 150 Mbps
$f_d = n \times f_o/c_o = (1{,}000$ m/s $\times$ 30 GHz)/3 $\times$ 108 m/s = 100 kHz
Differential Bit time $t_d = 1/150$ Mbps = 6.67 ns
Degree change (error) due to Doppler over t_d
$\Delta° = 100$ kcyc/s $\times$ 360°/cyc $\times$ 6.67 ns = 0.24° error
%error = 0.24°/180° $\times$ 100 = 0.13% for D-BPSK
%error = 0.24°/90° $\times$ 100 = 0.27% for D-QPSK, CP-PSK

These examples show that Doppler distortion for high-speed digital waveforms has minimal effect on the performance of the data link. Differential modulation is crucial in reducing the phase instabilities which improve the performance and decrease the cost for digital communications.

2.4.13 Minimum shift keying

MSK is also a constant envelope modulation with the only real difference from CP-PSK is that it is more of an analog version that is constantly changing and is not designed to stop at a phase location.

The sinusoidal weighting frequency is equal to half the bit rate, so that it eases the transition of a digital data bit by one cycle of the sine wave weighting signal. This provides a smoothing of the 180° phase shifts that occur in both of the channels. Then the phase transitions of the two channels are slower, which reduces the high-frequency content of the spectrum and results in attenuation of the sidelobes containing the high frequencies. Therefore, the spectrum is said to be efficient compared with standard PSK systems (e.g., BPSK, QPSK) since more power is contained in the main lobe and less in the sidelobes. The main lobe for MSK is 1.5 times larger than for OQPSK for the same number of bits sent due to the sinusoidal weighting.

Since the modulation is half the bit rate, the spectrum is widened by 1/2 bit width on both sides, which equals a 1-bit width bigger spectrum. Therefore, the total width would be three times the bit rate, which is 1.5(2 × bit rate). MSK is a continuous constant envelop modulation scheme, so it contains minimal AM and prevents sidelobe regeneration after filtering.

MSK can be generated in two ways. The first is sinusoidal weighting of the two BPSK channels in the OQPSK design before the summation takes place (Figure 2.33). This reduces the sidelobes and increases the main lobe. The other way to generate MSK is to use a FSK system, ensuring that the frequency separation is equal to the bit rate.

2.4.14 Frequency-shift keying

MSK can be generated using FSK by setting the frequency separation between the two frequencies equal to 1/2 the frequency shift rate or data rate. Therefore, MSK is a continuous phase FSK with a modulation index of 0.5. This is where MSK got its name, since it is the minimum spacing between the two frequencies that can be accomplished and still recover the signal with a given shift rate. It is rather remarkable that the same waveform can be produced via two different generation methods, each of which provides a different way of understanding, designing, and analyzing MSK systems. A simple way of generating MSK is shown in Figure 2.34.

A two-frequency synthesizer is frequency shifted by the binary bit stream according to the digital data. If the rate is set to the minimum bit rate, then MSK is the resultant output.

The frequency spacing of the FSK needs to be equal to 1/2 the bit rate to generate MSK. The frequency shift provides 1 bit of information, as does the BPSK waveform, but the null-to-null bandwidth is only 1.5 times the bit rate compared with 2 times the bit rate in the BPSK waveform. A simulation shows different types of FSK with different spacing with respect to the bit rate (Figure 2.35). If the frequency spacing is closer than 1/2 the bit rate, then the information cannot be recovered. If the spacing is too far apart, the information can be retrieved using

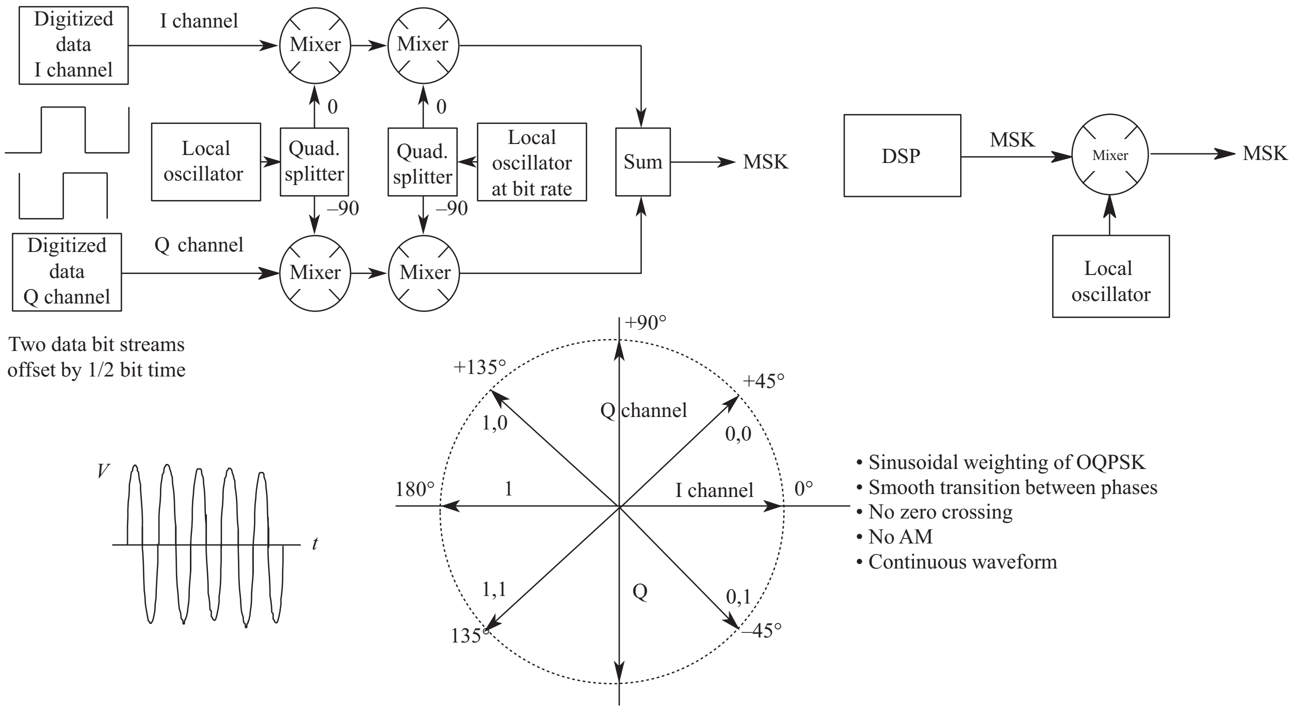


Figure 2.33 MSK generator

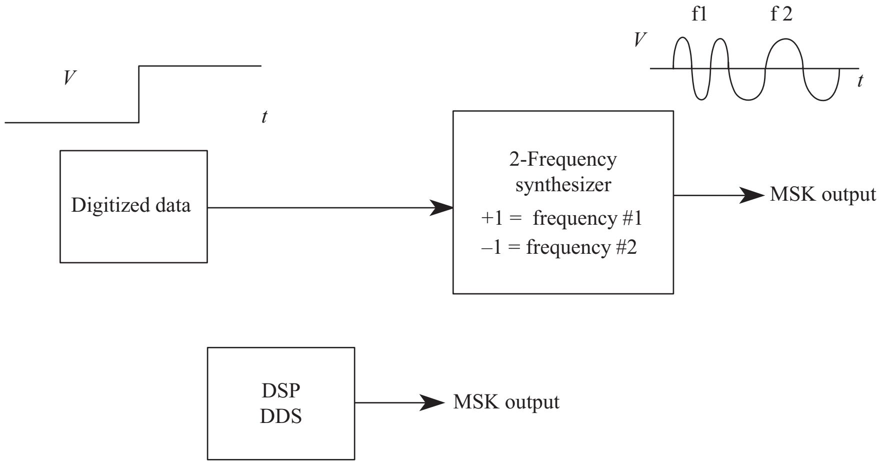


Figure 2.34 MSK generator using FSK

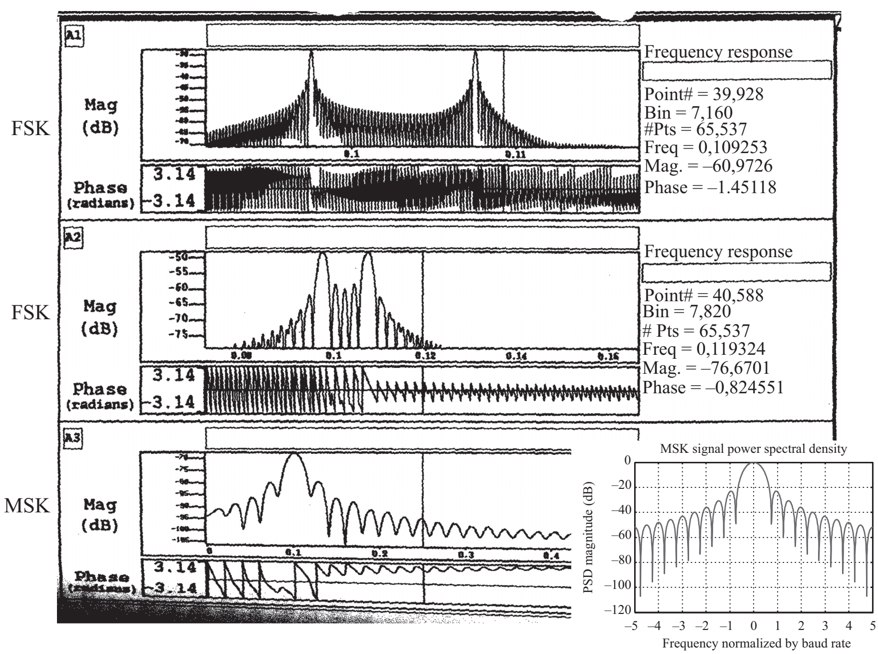


Figure 2.35 FSK spectrums with spacing minimized to generate MSK

FSK demodulation techniques, but MSK is not generated. As the frequency spacing approaches 1/2 the bit rate, then the resultant spectrum is MSK (Figure 2.35).

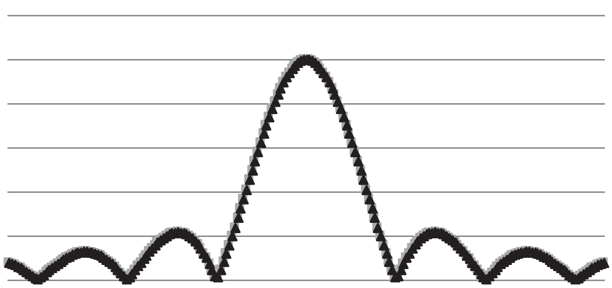
2.4.15 Sidelobe reduction methods

One of the problems with PSK systems is that the sidelobes can become fairly large and cause a problem with adjacent channel operation. The sidelobes continue out theoretically to infinity. The main concern is usually the first or second sidelobes, which are larger in magnitude. To confine the bandwidth for a particular waveform, a filter is required. The main problem with filtering a PSK signal is that this causes the waveform to be dispersed or spread out in time. This can cause distortion in the main signal and also more ISI, which is interference between the digital pulses.

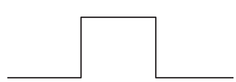
2.4.16 Ideal shaping filter

For no ISI, the sampling time of the pulse needs to occur when the magnitude of all other pulses in the digital signal is zero amplitude. A square wave impulse response in the time domain would be ideal, since there is no overlap of digital pulses, and thus, sampling anywhere on the pulse would not have ISI. However, the bandwidth in the frequency domain would be very wide, since a square wave in the time domain produces a sin(x)/x in the frequency domain.

Therefore, the best shaping filter to use for digital communications is a sin(x)/x impulse response in the time domain, which produces a square wave in the frequency domain. This provides ideal rejection in the frequency domain and also has a constant level with no AM and a sampling point in the center of the sin(x)/x *spectrum*. This produces the minimum ISI between symbols (Figure 2.36). However, a sin(x)/x impulse response is not possible since the sidelobes extend out to infinity. Thus, practical filter implementations are used to approximate this ideal



Impulse response filter in the time domain



Ideal frequency response of a time-domain impulse response sin (x)/x filter

Figure 2.36 Ideal sin(x)/x impulse response in the time domain

filter. Some of the practical filters used in communications are Gaussian, raised cosine, raised cosine squared, and root-raised cosine, which are all approximations of the $\sin(x)/x$ impulse response. The $\sin(x)/x$ impulse response approximates a square wave in the frequency domain, which reduces the noise bandwidth, reduces out-of-band transmissions, and is designed to minimize ISI (Figure 2.37).

Root-raised cosine filters are used extensively in communications and data links. They are employed in both the transmitter and the receiver, which basically splits the raised cosine filter, with half of the filter in the transmitter and half of the filter in the receiver. Therefore, the net result is a combination of the square root of the raised cosine filters, which leads to a raised cosine filter for the total data link system. The root-raised cosine filter has slightly faster transitions in the time domain for the transmitter pulse, and since a matched filter is used on the receiver side splitting the raised cosine filter response provides a slight improvement in the performance.

Another type of filtering scheme uses a Gaussian-shaped pulse. A Gaussian-shaped curve is a standard bell-shaped curve showing a Gaussian distribution, which is used extensively in probability theory (see Chapter 6). The Gaussian-shaped impulse response also provides a good approximation to the ideal $\sin(x)/x$ impulse response (Figure 2.37). This provides effective use of the band and allows multiple users to coexist in the same band with minimum interference. One type of modulation that includes this Gaussian impulse response is Gaussian MSK, a continuous phase modulation scheme that reduces the sidelobe energy of the transmitted spectrum. The main lobe is similar to MSK and is approximately 1.5 times wider than QPSK.

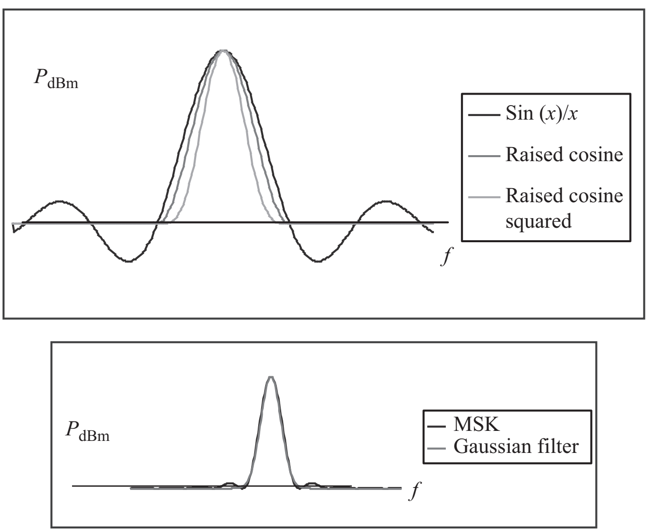


Figure 2.37 Practical digital waveform impulse responses

2.5 Direct sequence spread spectrum

Many systems today use spread spectrum techniques to transport data from the transmitter to the receiver. One of the most common forms of spread spectrum uses PSK and is referred to as direct sequence spread spectrum (DSSS). The data are exclusive-OR'd or modulo-2 added with a pseudo-random code that has a much higher chip rate than the data bit rate to produce the spread spectrum waveform. The bits in this high-speed PN code are called chips, to distinguish them from data bits. This produces a much wider occupied spectrum in the frequency domain (spread spectrum) (Figure 2.38).

DSSS systems use phase-shift generators (PSGs) to transfer data from the transmitter to the receiver by phase shifting a carrier frequency. This is done for both spread spectrum systems and nonspread spectrum systems. There are several ways to build a PSG depending on the waveform selected.

Fundamentally, spread spectrum uses more bandwidth than is required to send the signal. It utilizes techniques such as a faster pseudorandom code to spread the bandwidth. A higher code rate requires more bandwidth to send and receive the code.

2.5.1 Frequency-hopping spread spectrum

Most frequency-hopping spread spectrum transmitters use a direct frequency-hopping synthesizer for speed. Direct frequency-hopping synthesizers use a reference oscillator and directly multiply or translate the reference frequency to higher frequencies for use in the hopping sequence. An indirect frequency synthesizer uses a phase-locked loop (PLL) to generate the higher frequencies, which slows the process due to the loop bandwidth of the PLL. A PN code is used to determine which

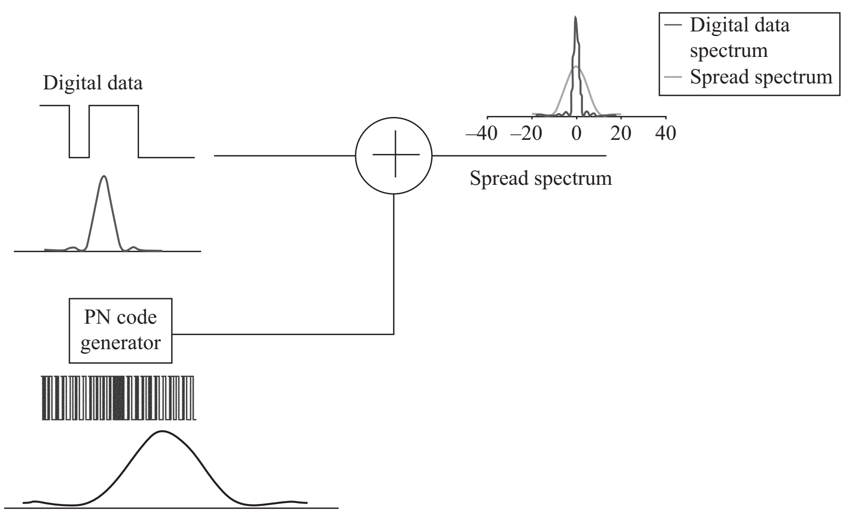


Figure 2.38 Direct sequence spread spectrum

of the frequency locations to hop to. The frequency hop pattern is therefore pseudorandom to provide the spread spectrum and the processing gain required for the system (Figure 2.39).

The process gain assumes that the jammer does not jam multiple frequency cells. Usually, the best jammer is a follower jammer. A follower jammer detects the frequency cell that the frequency hopper is currently in and immediately jams that frequency. The delay is associated with the time to detect the frequency and then to transmit the jammer on that frequency. For slow frequency-hopping systems, this method works quite well because most of the time the signal will be jammed. The faster the frequency hopping, the less effective the follower jammer is on the system. Moreover, many times, a combination of frequency hopping and direct sequence is used, slowing the detection process and reducing the effectiveness of the follower jammer.

2.5.2 Spread spectrum

The main reasons for using spread spectrum are as follows:

- It provides jammer resistance, reducing the effects of jamming.
- It allows multiple users to operate in a given band.
- It reduces interference with existing narrowband systems.
- It minimizes multipath effects; multipath is narrowband and dependent on frequency.
- It provides a LPI/LPD; the transmitted signals are hard to detect or intercept and are spread out so that they look like noise.

Spread spectrum was initially invented and patented in 1942 by an actress named Hedy Lamarr and a composer named George Antheil. It was used by the military during World War II for covert operations to transmit signals undetected. However, it was not fully utilized until many years later. Today, spread spectrum is commonplace and is used extensively in communications and in the development and design of data links. Two methods of spreading the spectrum is using a fast pseudorandom code with respect to the data rate or changing the frequency across a wideband (Figure 2.40).

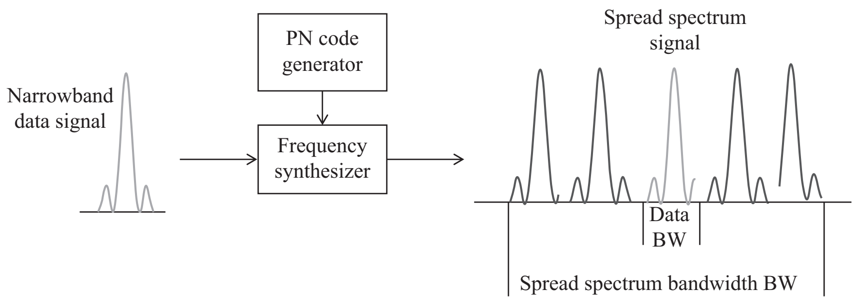


Figure 2.39 Frequency hopped spread spectrum

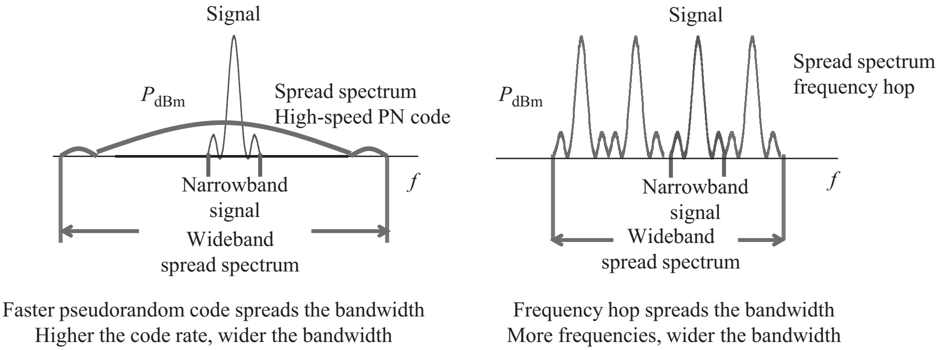


Figure 2.40 Two methods of spread spectrum use more bandwidth than is required to send data

2.5.3 Jammer resistance

One of the main advantages of spread spectrum systems is their ability to operate in a jamming environment, both friendly interfering signals and hostile jammers, where there is an attempt to jam communications. The more spreading that is used, the more jammer resistance the system has. An analysis that is used to determine a system's ability to resist jamming is called process gain (G_p). Process gain is essentially the spread bandwidth over the required bandwidth or how much band spreading the system is using.

The process gain for DSSS systems or PSK systems is:

$$G_p = 10\log(\text{RF bandwidth}/\text{detected bandwidth})$$

The process gain for frequency-hopping systems is equal to the occupied bandwidth over the signal bandwidth and is:

$$G_p = 10\log(\text{number of frequency channels})$$

The number of frequency channels assumes that they are independent and adjacent.

When calculating G_p for a frequency-hopping system, the separation of the frequencies used can alter the actual jamming margin. For example, if the frequencies are too close together, the jammer may jam multiple frequencies at a time. Moreover, the frequencies used need to be adjacent, with no missing frequencies, or the previous calculation becomes invalid.

The jamming margin (J_m), or the level of a jamming signal that will interfere with the desired signal, is equal to the process gain (G_p) minus the spreading losses (L_{ss}):

$$J_m = G_p - L_{ss}$$

This assumes that the signal-to-jammer ratio (SJR) is zero.

In addition, since most systems need a certain amount of signal level above the jamming level, the required SJR is specified. The required SJR will vary depending

on the modulation used and the bit error rate (BER) required. This needs to be included to calculate the actual jamming margin.

What follows is an example of calculating the jamming margin of a direct sequence system with a required SJR:

$$G_p = 10\log(1{,}000\ \text{MHz}/10\ \text{MHz}) = 20\ \text{dB}$$
$$J_m = 20\ \text{dB} - 3\ \text{dB} = 17\ \text{dB}$$
$$J_m(\text{actual}) = 17\ \text{dB} - \text{SJR(required)} = 17\ \text{dB} - 9.5\ \text{dB} = 7.5\ \text{dB}$$

A signal can be received at the specified BER with the jammer 7.5 dB higher than the desired signal.

One frequently asked question is, "Does spreading the signal increase the SNR of the system so that it requires less power to transmit?" In general, the spreading process does not improve the SNR for a system. Sometimes this can be a bit confusing when only looking at the receiver. The receiver by itself improves the SNR related to the input signal at the antenna. This improvement is included in the link budget because the bandwidth required for the spread signal is much larger on the input to the receiver than is needed for a nonspread system. Since the bandwidth is large on the input to the receiver, it is either reduced to lower the noise power, which produces a better SNR, or the signals are summed together, producing a larger amplitude and thus a better SNR.

However, if the entire spread spectrum system from the transmitter to the receiver is examined, the narrowband signal is not spread or despread and produces the same ideal SNR as the spread spectrum system. Therefore, in theory, the SNR required at the receiver for a nonspread system is the same as that required for a spread system.

In actual spread spectrum systems, however, the SNR is degraded due to the spreading losses associated with the spreading and despreading processes. Therefore, without jammers present and not considering possible multipath mitigation, it takes more power to send a signal using spread spectrum than it does to send the same signal using a nonspread spectrum system.

2.5.4 Despreading to realize process gain in the spread spectrum system

One of the ways to achieve the process gain in a spread spectrum system is to reduce the bandwidth, thereby lowering the noise. For example, a continuous BPSK signal uses a sliding correlator for despreading, which reduces the bandwidth. The bandwidth is spread due to the phase shift rate (chip rate) in the transmitter, and the receiver uses the same code and slides or lines up the code using a sliding correlator that multiplies the incoming signal with the chip code (Figure 2.41). This process strips off the high-speed code and leaves the slow data rate bit stream, which requires less bandwidth to process. The narrow bandwidth reduces the noise level, but the signal level is not increased. The jamming signal is spread out by the sliding correlator so that the amplitude of the jammer is lower than the desired signal in the narrow bandwidth which provides the process gain (Figure 2.41).

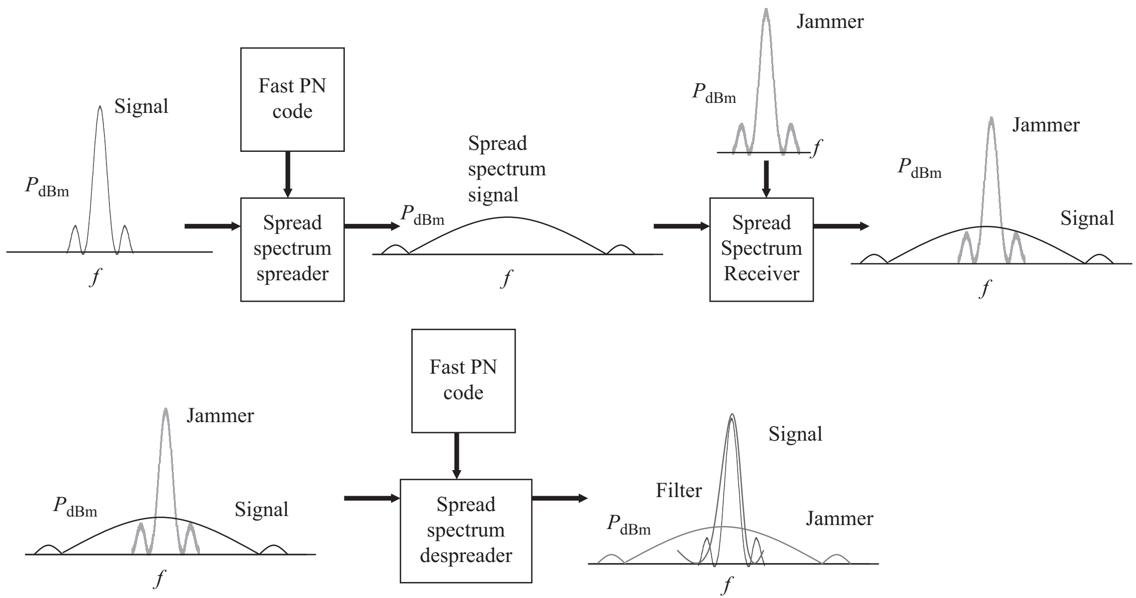


Figure 2.41 Direct sequence spread spectrum signal flow

Another example of reducing the bandwidth to achieve process gain is using a frequency-hopping signal that has constant amplitude but changes frequency over a large bandwidth at the transmitter. The dehopper in the receiver follows the hopped signal, and the resultant bandwidth is greatly reduced so that the noise is reduced by the same amount. However, the signal level is not increased, but the bandwidth is reduced.

The narrowband jammer is spread out by the dehopping process so that the average amplitude of the jammer is lower than the desired signal (Figure 2.42). The desired signal is jammed only for a short period of time as the jammer hops into the narrow bandwidth where the desired signal resides. The rest of the time, the frequency hopper is hopping to different frequencies outside the desired signal bandwidth, which gives the process gain against jammers. However, although the average power of the jammer is reduced because the energy is spread out over the bandwidth for an instant in time, the signal will be jammed while the jammer is in band. This may be critical in some systems where each hop frequency contains data that are lost for that short period of time. However, if the hop rate is higher than the data rate, then the jammer power can be looked at as a decrease in average power.

Another way to realize process gain is by summing the power in each bit of energy to increase the signal strength. This is accomplished in the analog domain by using a match filter correlator similar to an acoustic charge transport device or a surface acoustic wave (SAW) device. However, the preferred function is done in the digital domain by using a finite impulse response filter for the matched filter correlator with the tap (single delay in a shift register) spacing equal to the chip time, with the weights either $+1$ or -1 depending on the code pattern. These concepts will be discussed later. This digital system produces pulses when the code is lined up, and data can be extracted as the pulses are received or by using the time slot when the pulses occur for pulse position modulation. Although the bandwidth is not affected—and therefore the noise is not decreased—it does integrate the signal up to a larger level. The process gain is 10 log (number of bits combined). The jammer is not coherent to the matched filter correlator, so it does not correlate with the code. However it does produce some lower amplitude correlation for short codes which can be further reduced by using a longer code and a longer matched filter correlator.

All of these approaches are used for LPI systems since the signal is spread out and hard to detect unless a matched receiver with the known spread spectrum characteristics are known.

The main reasons for using spread spectrum are to reduce the effects of jamming and interfering signals, to reduce the ability for detection by an unwanted receiver, and to facilitate multipath mitigation. However, remember that it takes more power to send a spread spectrum signal due to spreading losses in the presence of noise only.

2.5.5 Maximal length sequence codes

Maximal length sequence (MLS) codes, referred to as m-sequence codes, are used extensively in digital communication systems. The reason is that they have very

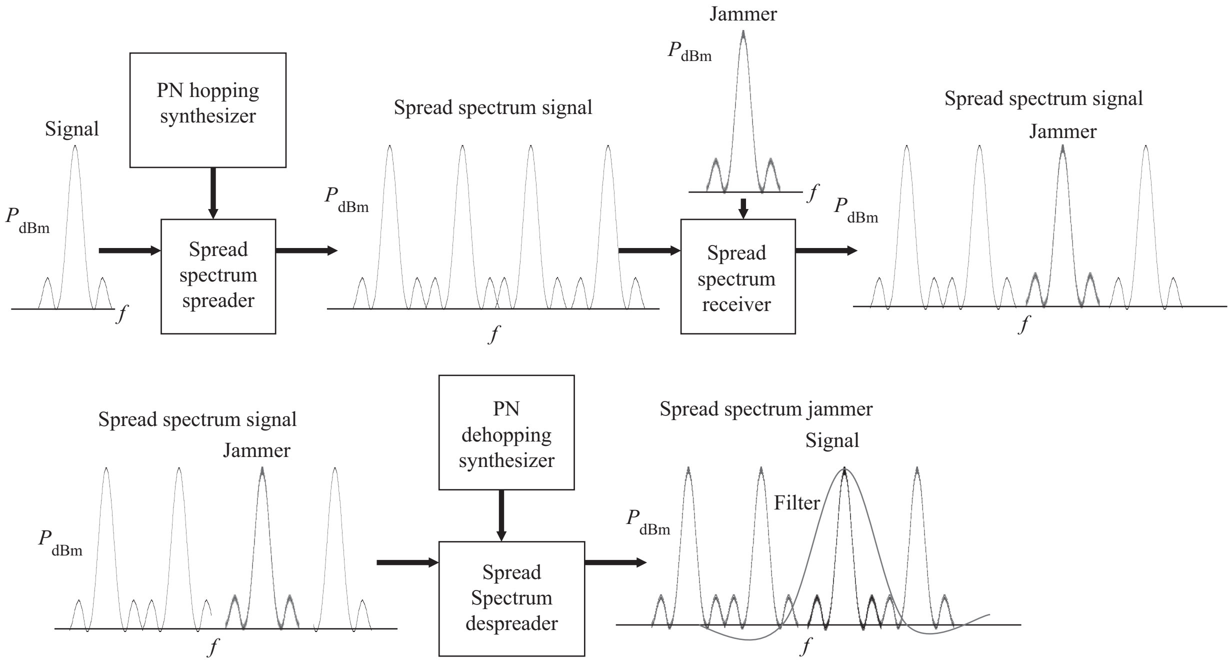


Figure 2.42 Frequency hopped spread spectrum

good properties for minimizing interference between codes by reducing the cross-correlation products or the amount of correlation with the wrong code.

Some of the code properties that they contain are the following:

- Runs (sections of code) are defined: Half of the runs are of length 1, one-quarter of the runs are of length 2, one-eighth of the runs are of length 3. The length is the amount of consecutive "1"s or "0"s in a code.
- Contains one more "1" as compared with the number of "0"s.
- Their autocorrelation function is 1 for zero delay, $-1/(2^m - 1)$ for any delay (m is the number of shift registers and $2^m - 1$ is the length of the code), positive or negative, greater than 1 bit with a triangular shape between ± 1 bit.
- Generated by a linear feedback shift register (LFSR).
- The taps of the LFSR are defined depending on the polynomial.
- Tap numbers will all be relatively prime.

These codes, or a modification of these codes, are used in nearly all digital communication systems. In the case of the global positioning system, the different m-sequence codes for the different satellites are generated by using an m-sequence code and by multiplying this code with different delays of the same code.

2.5.6 Maximal length PN code generator

A method of building a PN generator consists of a shift register and a modulo-2 adder. The taps (digital delays) used from the shift register to the modulo-2 adder are specified for each of the shift register lengths to generate a MLS code. If a MLS code can be chosen so that only two taps are required, a simple two-input exclusive-or gate can be used. A "1" is loaded into the shift registers on power-up to get the code started (Figure 2.43).

A counter is provided to ensure that the "all-zeros case" is detected (which would stop the process), and a "1" is automatically loaded into the PN generator. The basic building block for generating direct sequence PSK systems is the MLS generator. There are other methods of generating MLS codes. One process is to store the code in memory and then recall the code serially to generate the MLS code stream. This provides the flexibility to alter the code, for example, to make a perfectly balanced code of the same number of "0"s as "1"s (orthogonal code), which is a modification to the MLS generator. This provides a code with no direct current (DC) offset. The DC offset in the system causes the carrier to be less suppressed because of the bias. This added bias causes the carrier to be present in a suppressed carrier system.

Other types of codes can be generated, such as the Gold codes used to provide codes that reduce the crosscorrelation between different code sets.

2.5.7 Maximal length PN code taps

To generate a maximal length code using the code generator previously discussed, the taps of the shift register are defined by a maximal length tap table. As the length

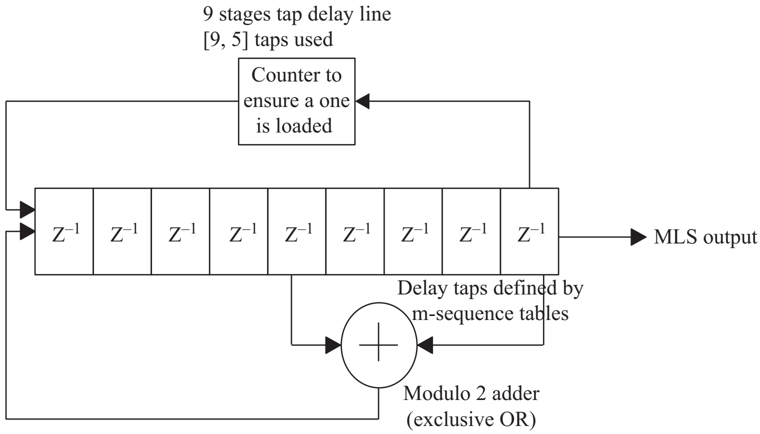


Figure 2.43 Maximal length sequence code generator

Table 2.1 Maximally length PN code taps

```
3 stages: [3, 2]
4 stages: [4, 3]
5 stages: [5, 3][5, 4, 3, 2][5, 4, 3, 1]
6 stages:[6, 5][6, 5, 4, 1][6, 5, 3, 2]
7 stages:[7, 6][7, 4][7, 6, 5, 4][7, 6, 5, 2][7, 6, 4, 2][7, 6, 4,
1][7, 5, 4, 3][7, 6, 5, 4, 3, 2][7, 6, 5, 4, 2, 1]
8 stages:[8, 7, 6, 1][8, 7, 5, 3][8, 7, 3, 2][8, 6, 5, 4][8, 6, 5,
3][8, 6, 5, 2][8, 7, 6, 5, 4, 2][8, 7, 6, 5, 2, 1]
9 stages:[9, 5][9, 8, 7, 2][9, 8, 6, 5][9, 8, 5, 4][9, 8, 5, 1][9, 8,
4, 2][9, 7, 6, 4][9, 7, 5, 2][9, 6, 5, 3][9, 8, 7, 6, 5, 3][9, 8, 7, 6,
5, 1][9, 8, 7, 6, 4, 3][9, 8, 7, 6, 4, 2][9, 8, 7, 6, 3, 2][9, 8, 7, 6,
3, 1][9, 8, 7, 6, 2, 1][9, 8, 7, 5, 4, 3][9, 8, 7, 5, 4, 2][9, 8, 6, 5,
4, 1][9, 8, 6, 5, 3, 2][9, 8, 6, 5, 3, 1][9, 7, 6, 5, 4, 3][9, 7, 6, 5,
4, 2][9, 8, 7, 6, 5, 4, 3, 1]
10 stages:[10, 7][10, 9, 8, 5][10, 9, 7, 6][10, 9, 7, 3][10, 9, 6,
1][10, 9, 5, 2][10, 9, 4, 2][10, 8, 7, 5][10, 8, 7, 2][10, 8, 5, 4][10,
8, 4, 3][10, 9, 8, 7, 5, 4][10, 9, 8, 7, 4, 1][10, 9, 8, 7, 3, 2][10,
9, 8, 6, 5, 1][10, 9, 8, 6, 4, 3][10, 9, 8, 6, 4, 2][10, 9, 8, 6, 3,
2][10, 9, 8, 6, 2, 1][10, 9, 8, 5, 4, 3][10, 9, 8, 4, 3, 2][10, 9, 7,
6, 4, 1][10, 9, 7, 5, 4, 2][10, 9, 6, 5, 4, 3][10, 8, 7, 6, 5, 2][10,
9, 8, 7, 6, 5, 4, 3][10, 9, 8, 7, 6, 5, 4, 1][10, 9, 8, 7, 6, 4, 3,
1][10, 9, 8, 6, 5, 4, 3, 2][10, 9, 7, 6, 5, 4, 3, 2]
```

of the shift register increases, the number of possible tap configurations increases. A list of tap values is shown in Table 2.1.

2.5.8 Gold codes

Gold codes are a special case of maximal length codes. Simply put, they are generated by linearly combining two MLS codes and selecting only certain pairs, known as

preferred pairs to provide minimum crosscorrelation properties. An extension of the Gold codes is making the code orthogonal by adding a zero to the Gold code.

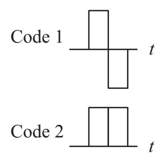
2.5.9 Other codes

Many other codes are used in digital communication, some of which are as follows:

- Kasami sequences: These codes are used and also have low crosscorrelation properties.
- Orthogonal codes: The Hadamard transform generates these codes, which, if they are ideal, have no crosscorrelation. However, if they are nonideal, any offset causes problems with crosscorrelation.
- Walsh codes: These codes are generated from an orthogonal set of codes defined using a Hadamard matrix of order *N*. The generator block outputs a row of the Hadamard matrix specified by the Walsh code index.

In this simple example of a Walsh orthogonal code, first of all, the number of bits equals the number of users. Therefore, 2 bits = two users. Since the codes are orthogonal:

Code 1 = (1,−1) Code 2 = (1,1) Orthogonal



$\langle X_1, X_2 \rangle = \int x_1(t) \times x_2(t)dt = 0$ Orthogonal
1,−1 × 1,1 = 1+−1 = 0 Orthogonal

Incorporating data with the codes we have:

Code 1 = (1,−1) Data 1 = **1011** (0 = −1)
Code 2 = (1,1) Data 2 = **0011** (0 = −1)
Bit Stream 1 = (1,−1) × (1,−1,1,1) = (1,−1,−1,1,1,−1,1,−1)
Bit Stream 2 = (1,1) × (−1,−1,1,1) = (−1,−1,−1,−1,1,1,1,1)
(1,−1,−1,1,1,−1,1,−1) + (−1,−1,−1,−1,1,1,1,1) = (0,−2,−2,0,2,0,2,0) Orthogonal
Decode Data 1: (1,−1) × (0,−2,−2,0,2,0,2,0) = (1,−1) × [(0,−2)(−2,0)(2,0)2,0)] = (0+2)(−2+0)(2+0)(2+0) = (2,−2,2,2) = **1,0,1,1** = Data 1
Decode Data 2: (1,1) × (0,−2,−2,0,2,0,2,0) = (1,1) × [(0,−2)(−2,0)(2,0)2,0)] = (0−2)(−2+0)(2+0)(2+0) = (−2,−2,2,2) = **0,0,1,1** = Data 2

2.5.10 Spectral lines in the frequency domain

Another criterion to be considered when selecting codes for communications is the characteristics of the spectral lines (unwanted frequencies) generated in the frequency domain. This is especially important for short codes, which will produce

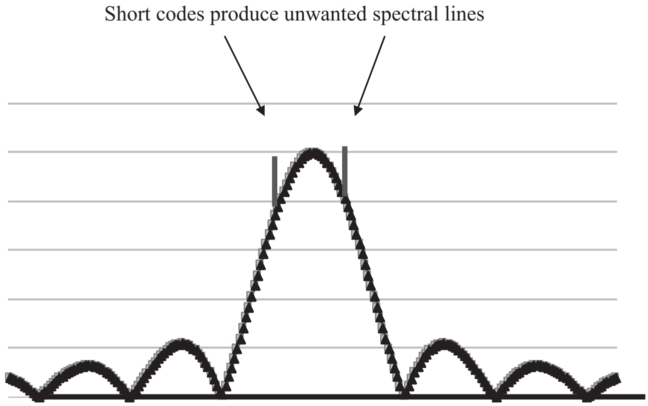


Figure 2.44 Spectral lines caused by repetitions in the code

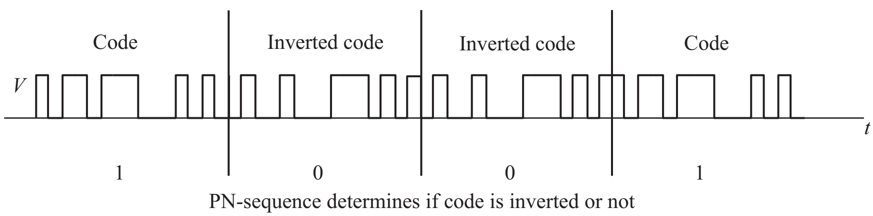


Figure 2.45 Invert the short code using a pseudorandom sequence to reduce spectral lines

spectral lines caused by the code repetition frequency and sections of the code that repeat (Figure 2.44). Basically, anything that repeats in the time domain will show up as a spectral frequency in the frequency domain. This can pose a problem with requirements for power spectral density (frequency content) in the specified band. One method of reducing these spectral lines in the frequency domain is to invert the existing PN-code according to an additional PN-code. This additional PN-code determines when to invert or not invert the existing PN-code (Figure 2.45). This technique has been patented and has been proven successful in reducing the spectral lines to an acceptable level.

2.6 Other forms of spread spectrum transmissions

Other forms of spread spectrum include time hopping and chirped-FM. The time-hopping system uses the time of transmission of the data and ignores the rest of the time, which reduces the overall effect of the jammer that has to be on for the entire

time. The chirped system generates more bandwidth by sweeping over a broad bandwidth. This system reduces the bandwidth in the detection process by reversing the process, thus providing process gain.

2.6.1 Time hopping

Time hopping entails transmitting the signal only at specified times, that is, the transmissions are periodic, and the times to transmit are pseudorandom using a pseudorandom code (Figure 2.46). The process gain is equal to 1/duty cycle. This means that if the duty cycle is short—on for a short period of time and off for a long period of time—the process gain is high.

The receiver demodulates this by looking only at the times that the signal was sent, knowing the pseudorandom sequence of time slots. This increases the level of the signal during the time slots and decreases the average power of the jamming signal by ignoring the other time slots since the jamming signal is spread over all the pulses.

2.6.2 Chirped-FM

Chirped-FM signals are generated by sliding the frequency in a continuously changing manner, similar to sweeping the frequency in one direction. The reason that these signals are called chirped is that when spectrally shifted to audible frequencies, they sound like a bird chirping. There are up-chirps and down-chirps, depending on whether the frequencies are swept up or down. Chirp signals can be generated by a sweeping generator with a control signal that resets the generator to the starting frequency at the end of the chirp. This is generally a slow process and is not as effective as a higher speed sweep. Chirp signals can also be generated using SAW devices excited by an impulse response (which theoretically contains all frequencies). The delay devices or fingers of the SAW device propagate each of the frequencies at different delays, thus producing a swept frequency output. This reduces the size of chirping hardware tremendously and produces a very fast sweep time.

For chirp-FM systems, the instantaneous frequency is calculated by

$$F = F_h - t(W/T)$$

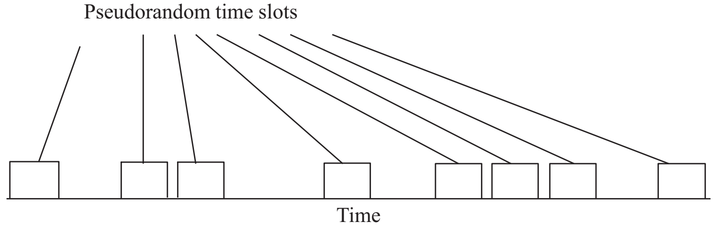


Figure 2.46 Time hopping spread spectrum signal

where F is the instantaneous frequency, F_h is the highest frequency in the chirped bandwidth, t is the change in time (Δt), W is the bandwidth of the dispersive delay line (DDL), and T is the dispersive delay time.

For chirp systems, the high frequencies have more attenuation for a given delay. For a flat amplitude response for all frequencies, the low frequencies need to be attenuated to match the high frequencies. Therefore, the total process contains more insertion loss. SAW devices are used to provide frequency-dependent delays, since it is easier to control the delay with lower frequency sound waves than electromagnetic waves. These are called acoustic delay lines or delay fingers.

The up-chirp is used by the transmitter; the down-chirp is used by the receiver, matched to the transmitter to eliminate the chirp (Figure 2.47). The up-chirp spreads the frequencies over a broad band, and the down-chirp despreads the signal. In addition, the jammer, when in the down-chirp, since it is not spread by the up-chirp, is spread across a broad band, thus reducing the impact that the jamming signal has on the system.

2.7 Multiple users

There are three basic methods that separate multiple users in the same geographical area which prevents interference between the uses. They are as follows:

- Time division multiple access (TDMA)
- Code division multiple access (CDMA)
- Frequency division multiple access (FDMA)

These techniques are shown in Figure 2.48. Some control is required in all of these systems to ensure that each system in an operating area has different assignments.

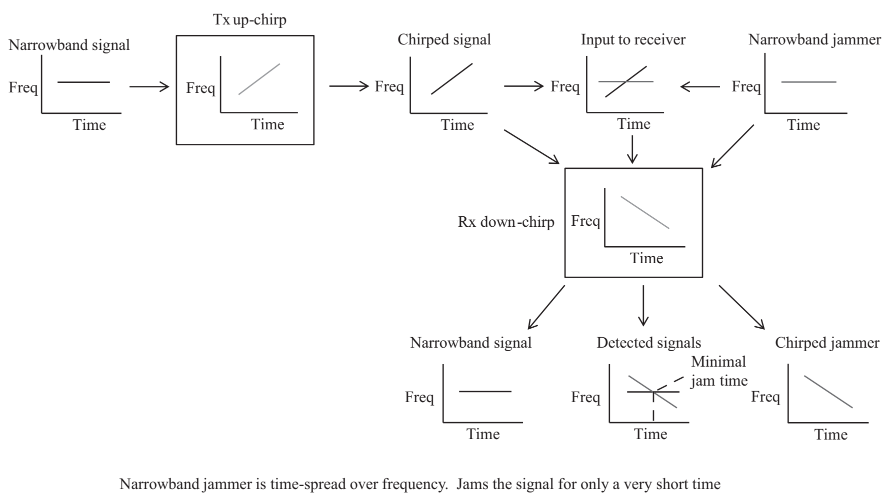


Figure 2.47 Chirped FM spread spectrum

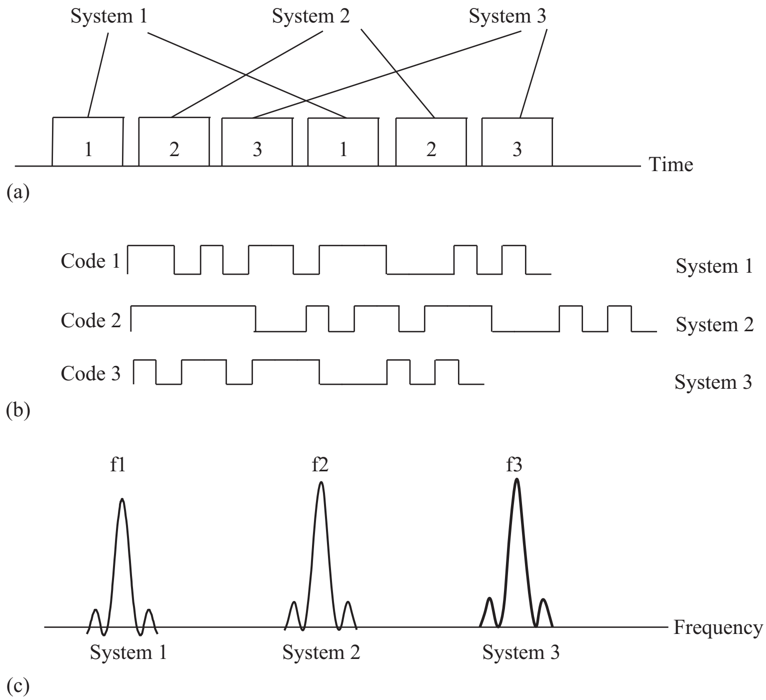


Figure 2.48 Multiple user techniques applying TDMA, CDMA, and FDMA: (a) time division multiple access, (b) code division multiple access, and (c) frequency division multiple access

TDMA provides interference reduction by having the systems communicate at different times or time slots using the same frequencies and codes (Figure 2.48). If the systems have predetermined time slots, then the system is considered to be time division multiplexing (TDM). If a system accesses the time slot, say, on a first-come, first-serve basis, then the system is a TDMA system. This definition applies to other types of systems, such as CDMA and code division multiplexing, and FDMA and frequency division multiplexing (FDM).

CDMA provides interference reduction by having the systems communicate on different codes, preferably MLS codes, orthogonal codes, or Gold codes, at the same frequencies and times, which provide minimum crosscorrelation between the codes, resulting in minimum interference between the systems (Figure 2.48). A lot of research has been done to find the best code sets for these criteria. Generally, the shorter the code, the more crosscorrelation interference is present, and fewer optimal codes can be obtained.

FDMA provides interference reduction by having the systems communicate on different frequencies, possibly using the same times and codes (Figure 2.48). This provides very good user separation since filters with very steep roll-offs can be

used. Each user has a different frequency of operation and can communicate continuously on that frequency.

Each of the previously discussed multiuser scenarios reduces interference and increases the communications capability in the same geographical area.

2.7.1 Other methods for multiuser techniques

Other methods to allow multiple users in the same band of frequencies include frequency hopping, spatial separation, orthogonal techniques, and wideband–narrowband signal separation. Frequency hopping is where each user has a different hop pattern. Spatial separation is where each user is in a different area, and they are separated in space or direction (Figure 2.49). Pointing directional antennas to different sectors allows more users to operate in the same band. Orthogonal techniques are used for separating users where each user is orthogonal to the others in the band. Finally, narrowband and wideband users can be separated using correlation techniques. Wideband signals do not have autocorrelation with a chip/bit delay, whereas narrowband signals have autocorrelation with the same delay. This technique will be utilized in the adaptive filter discussion to mitigate narrowband jammers in a wideband signal environment.

2.7.2 Orthogonal signals

Several types of orthogonal signals can be used to separate multiple communication links in the same band. This is valuable for multiple access applications and for separation of channels to reduce adjacent channel interference. Some of the techniques used to separate signals by orthogonal means include the following:

- Phase separation: separating I/Q or cosine and sine signals.
- Orthogonal frequency separation: OFDM, which uses orthogonal frequencies.

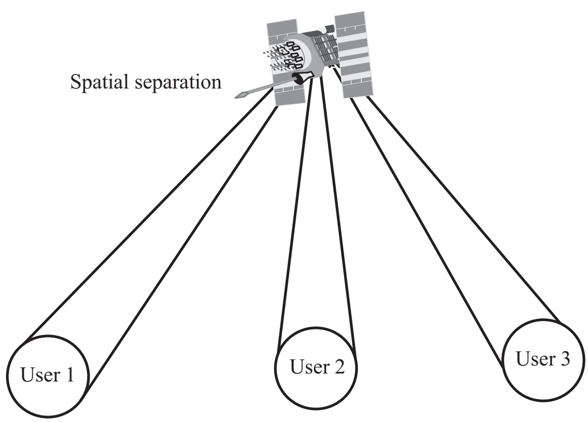


Figure 2.49 Directional beam antennas provide spatial separation for multiple users

- Antenna polarization: separation due to vertical versus horizontal, left-hand circular polarized versus right-hand circular polarized.
- Jammer/signal separation: Graham-Schmidt orthogonalizer forces the jammer to be orthogonal to the signal using error feedback.

2.7.3 Quadrature phase detection of two signals

Quadrature phase detection is the ability to separate signals due to orthogonal phases. For example, if one signal can be put on an in-phase carrier, $\cos(\omega t)$, and another signal put on an orthogonal signal, 90° out of phase of the first signal, $\sin(\omega t)$, then these signals can be separated by the receiver (Figure 2.50).

A simple math calculation shows that the two signals will be on each of the paths, as shown as follows:

- Path 1: Low-pass filter included to eliminate the sum term.
    - f_1 signal: $f_1(t)\cos(\omega t) \times \cos(\omega t) = f_1(t)[1/2\cos(\omega t - \omega t) + 1/2\cos(\omega t + \omega t)] = 1/2 f_1(t)[\cos(0)] = 1/2 f_1(t)$; signal f_1 is detected.
    - f_2 signal: $f_2(t)\sin(\omega t) \times \cos(\omega t) = f_2(t)[1/2\sin(\omega t - \omega t) + 1/2\sin(\omega t + \omega t)] = 1/2 f_2(t)[\sin(0)] = 0$; signal f_2 is not detected.
- Path 2: Low-pass filter included to eliminate the sum term.
    - f_1 signal: $f_1(t)\cos(\omega t) \times \sin(\omega t) = f_1(t)[1/2\sin(\omega t - \omega t) + 1/2\sin(\omega t + \omega t)] = 1/2 f_1(t)[\sin(0)] = 0$; signal f_1 is not detected.
    - f_2 signal: $f_2(t)\sin(\omega t) \times \sin(\omega t) = f_2(t)[1/2\cos(\omega t - \omega t) - 1/2\cos(\omega t + \omega t)] = 1/2 f_2(t)[\cos(0)] = 1/2 f_2(t)$; signal f_2 is detected.

Therefore, both the signals will be retrieved, with signal f_1 appearing on path 1 and signal f_2 appearing on path 2. A phasor diagram shows the separation of the two signals (Figure 2.50).

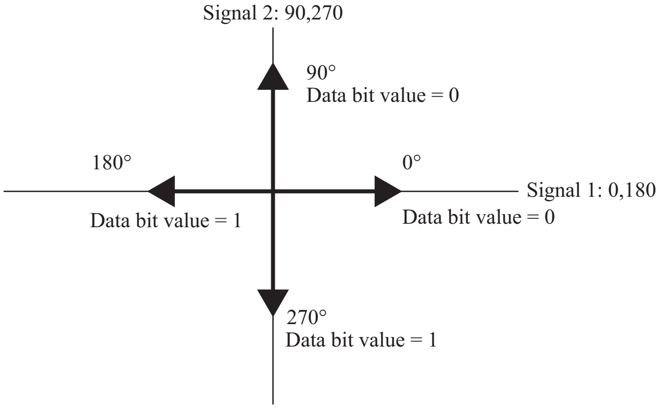


Figure 2.50 Quadrature signals

A practical detector called a Costas loop uses feedback to phase lock on a carrier frequency to provide two phases at the output of the detection process (Figure 2.51). Costas loops will be discussed in detail later in this book.

2.7.4 Orthogonal frequency division multiplexing

An extension of FDM uses the principle that if signals are orthogonal, they do not interfere with each other. A technique being used in many communications applications is known as OFDM. This technique allows closer spacing of multiple users in a given bandwidth or allows for multiple parallel channels to be used for higher data rates by combining parallel channels. It is used for broadband communications in both powerline communications and RF systems with different modulation schemes, including BPSK, QPSK, and QAM. The basic concept is to use frequencies that are orthogonal over a specified time period.

The basic orthogonality concept is that the inner product of two orthogonal frequency channels are integrated to zero over a specified integration time and the inner product of the same frequency is equal to one:

$$\langle Xn(t), Xm(t)\rangle = \int_0^t Xn(t)Xm(t)dt = 1$$

where $Xn(t) = Xm(t)$ and

$$\langle Xn(t), Xm(t)\rangle = \int_0^t Xn(t)Xm(t)dt = 0$$

where $Xn(t), Xm(t)$ are orthogonal

Consequently, if the frequencies are orthogonal, their inner products are equal to zero. So taking the inner product of all the frequencies with the desired frequency, only the incoming desired frequency will be present.

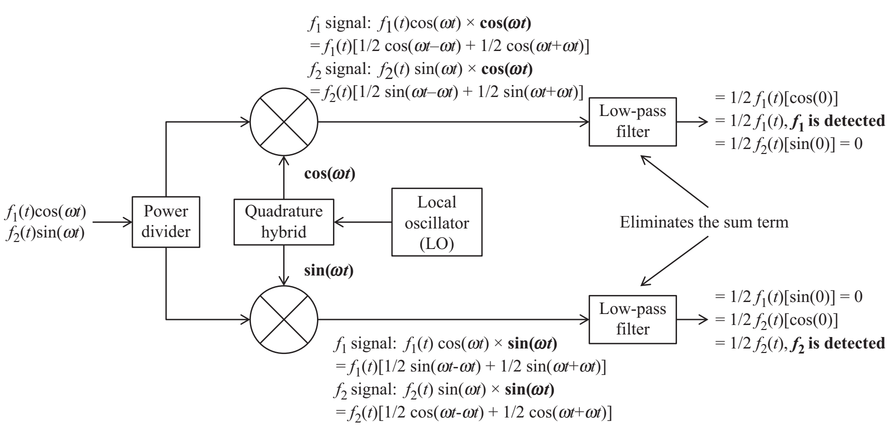


Figure 2.51 Quadrature detection using orthogonal phase separation

One way to obtain orthogonal frequencies is to ensure that the frequencies are harmonics of the fundamental frequency. Then, by choosing the integration time, the inner product will be zero. An example of orthogonal frequencies containing the second and third harmonic is shown in Figure 2.52.

When any combination of these frequencies is multiplied and integrated with the limits shown in Figure 2.52, the results are equal to zero. If the inner product of any one of these frequencies with itself is used, then the inner product is not zero, and the specified signal is detected.

A way to visualize this is to take the previous example, multiply every point on the curves, and sum the values. They will be equal to zero for the orthogonal frequencies and equal to a one for the same frequencies. A simple example was shown earlier using $\cos(\omega t)$ and $\sin(\omega t)$ as the orthogonal signals.

For OFDM, there are multiple users for each of the frequency slots, and the adjacent slots are made to be orthogonal to each other so that they can overlap and not interfere with each other. This allows more users to operate in the same bandwidth. Thus, using orthogonal frequencies reduces adjacent channel interference and allows for more channels to be used is a smaller bandwidth (Figure 2.53).

Another use for OFDM is to allow high-speed data to be transmitted in a parallel system. This is accomplished by taking the serial data stream from the transmitter and converting to parallel streams, which are multiplexed using different frequency channels in the OFDM system to send the parallel stream of data to the receiver. The receiver then performs a parallel-to-serial conversion to recover the data. In addition, the system can be made adaptive to only select the channels that are clear for transmission.

One of the drawbacks to OFDM is the high CF (Figure 2.3). The CF or PAPR is greater than 10 dB. This requires a linear PA, 10 dB from saturation, and a high dynamic range of A/D and D/A. Due to several waveforms with high CFs, there have been several PAPR reduction techniques to help mitigate these problems as follows:

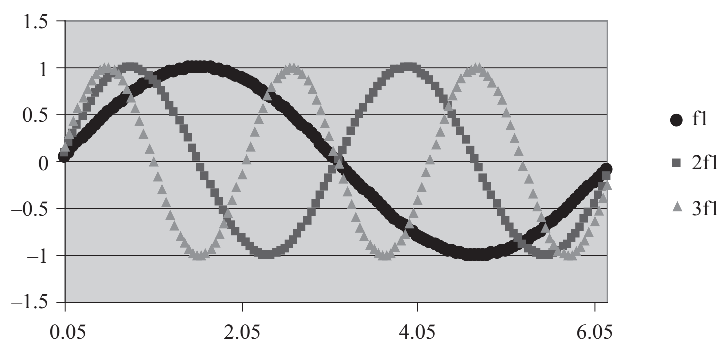


Figure 2.52 Orthogonal frequencies producing an inner product of zero

1. Use of Golay or Reed/Muler codes help to reduce the PAPR.
2. Several CF reduction techniques have been implemented for multiple signals used in long-term evolution (LTE) applications.
3. Use of digital predistortion can help to reduce the CF.

2.7.5 Other OFDM techniques

To improve the performance of a standard OFDM system, additional modulation techniques have been added. For example, two types of OFDM systems that have aided other techniques in performance improvement are as follows:

- Coded OFDM: implements coding schemes in addition to orthogonal techniques to improve the performance against multipath.
- Vector OFDM: uses TDM and packet communications to improve wireless broadband networks, including local multipoint distribution service systems.

2.8 Power control

Power control is a technique to control the amount of power of multiple transmitters into a single receiver. This is especially important for systems using CDMA and other multiplexing schemes. Power control helps reduce the effects of near/far problems where one user is close to the transceiver and another user is far away. The user close to the receiver will jam the user that is far away from the receiver. If power control is used, the idea is that the power will be the same level at the receiver for both the close-in and far-away users.

A multiuser system or network is based on the premise that the users are all at the same power level, and the separation between users is accomplished by

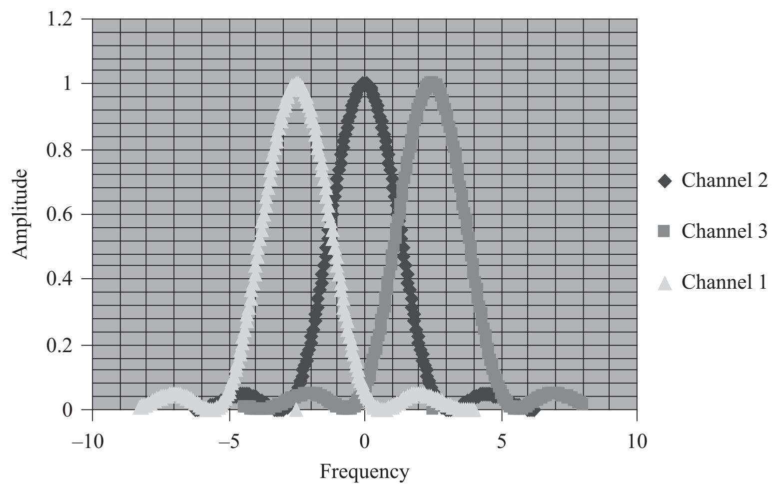


Figure 2.53 OFDM allows frequency channels to overlap with minimal adjacent channel interference

multiplexing schemes such as CDMA. However, most users are at different locations and distances from the receiver so that the power into the receiver varies tremendously, up to 100 dB. Since most CDMA systems can handle variations in power or jamming levels, there still exist problems with high level interference from other users.

Power control helps to mitigate this interference problem by having the user transmit only the amount of power required for the distance from the receiver. Therefore, for a user close to the receiver, the power is reduced. And for a user at a far distance from the receiver, the power is increased. The main objective is to have the power, regardless of the distance, at the same level when it reaches the receiver (Figure 2.54).

Many digital cell phone networks use power control to allow more users and less interference. The base station controls the power of all of the telephones that are communicating with that particular base station so that the power is nearly the same regardless of the distance from the base station. With the limited amount of process gain and jamming margin, a typical CDMA system is required to have power control to operate correctly. Power control is used where process gain is inadequate for good separation of users.

2.9 Summary

The transmitter is a key element in the design of the transceiver. The transmitter provides a means of sending out the information, over the channel, with the power

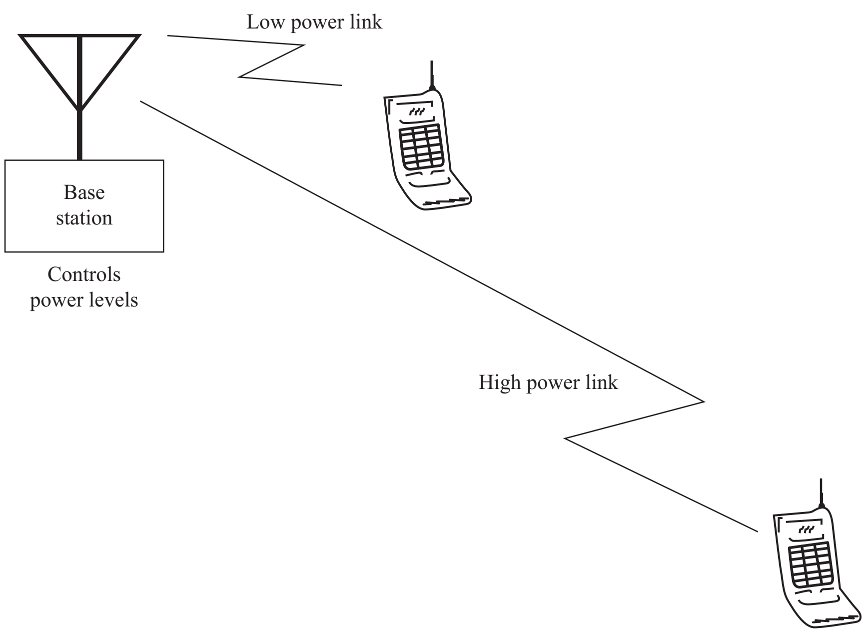


Figure 2.54 Power control

necessary to provide coverage to the intended receiver. Several types of modulation methods were discussed with their advantages and disadvantages including CP-PSK which is an optimal waveform for communications including pulsed systems and saturation. Differential modulation is incorporated to help mitigate Doppler, phase noise from oscillators, and other disturbances. CF, sum and difference products, up/down converters, SDRs, and their usage are discussed in this chapter. Many types of spread spectrum transmitters provide process gain to reduce the effects of jammers and to allow more efficient use of the spectrum for multiple users. Digital systems have many advantages over analog systems, and different techniques were discussed for optimizing the digital data link.

2.10 Problems

1. What is the Crest Factor and why is it important for required power?
2. What do we make the load impedance equal to for maximum power transfer?
3. What is the main advantage of a digital data link over an analog data link?
4. What is a software-defined radio (SDR)? What is a cognitive radio?
5. What is BPSK? What is 16-QAM? What are their advantages/disadvantages?
6. What does EVM mean, and what does it measure?
7. What is the advantage of an MSK or CP-PSK over standard BPSK and QPSK systems?
8. Which two approaches can be used to generate MSK?
9. What types of filters are used to approach an ideal match filter? What is the result in the frequency domain of an ideal match filter in the time domain?
10. What is spread spectrum?
11. What are the reasons to use spread spectrum? What would be the reason to not use spread spectrum?
12. What are two basic techniques used to generate spread spectrum?
13. What is the basic concept of process gain with respect to bandwidths?
14. What devices can be used to generate spread spectrum codes?
15. What causes unwanted spectral lines in a modulated spectrum?
16. What techniques are used to remove them?
17. What are the three main techniques that can be incorporated to allow multiple users?
18. What are orthogonal codes used for?
19. What are the advantages and disadvantages of differential systems versus coherent systems?
20. What is OFDM? How does it increase the number of users in a bandwidth?
21. Why is power control needed for a CDMA data link?
22. Which two factors are needed to calculate the jamming margin from process gain?
23. Which technique does direct sequence use to generate spread spectrum?

Further reading

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Chapter 3

The receiver

The receiver is responsible for downconverting, demodulating, decoding, and unformatting the data received over the link with the required sensitivity and bit error rate (BER) according to the link budget analysis of Chapter 1. The receiver is responsible for providing the dynamic range (DR) to cover the expected range and power variations and to prevent saturation from larger power inputs and provide the sensitivity for low-level signals. The receiver provides detection and synchronization of the incoming signals to retrieve the data sent by the transmitter. The receiver section is also responsible for despreading the signal when spread spectrum signals are used.

The main purpose of the receiver is to take the smallest input signal, the minimum detectable signal (MDS), at the input of the receiver and amplify that signal to the smallest detection level at the analog-to-digital converter (ADC) while maintaining a maximum possible signal-to-noise ratio (SNR). A typical block diagram of a receiver is shown in Figure 3.1. Each of the blocks will be discussed in more detail.

3.1 Superheterodyne receiver

Heterodyne means to mix, and superheterodyne means to mix twice. Most receivers are superheterodyne receivers, which means that they use a common intermediate frequency (IF) and two stages or double conversion to convert the signal to a lower frequency or baseband, as shown in Figure 3.1. The reasons for using this type of receiver include the following:

- A common IF can be used to reduce the cost of parts and simplify design for different radio frequency (RF) operating ranges.
- It is easier to filter the image frequency since the image frequency is far away at two times the IF.
- It is easier to filter intermodulation products (intermods) and spurious responses.
- Used in tunable RF frequencies and FH signals
  - Synthesizer changes frequencies but the IF filter and circuitry remain the same
  - It has the ability to change or hop the RF local oscillator (LO) for a constant IF chain

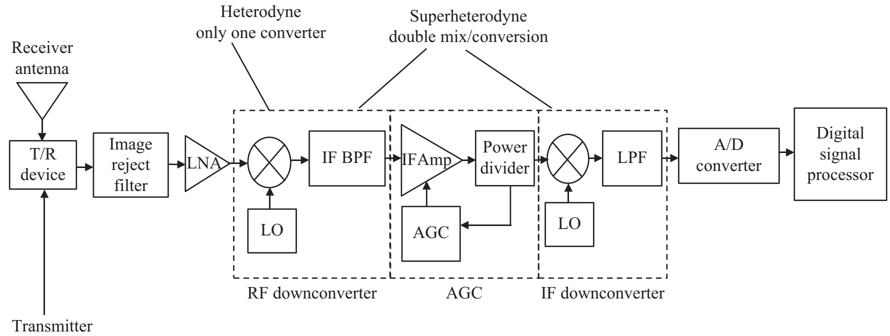


Figure 3.1 Typical superheterodyne receiver

Multiple-stage downconverters with more than one IF are sometimes used to aid in filtering and commonality. Occasionally, a single downconverter is used, and it generally contains an image-reject mixer to reduce the effects of the image frequency.

3.2 Basic functions of the receiver

The basic functions of the receiver are shown in Figure 3.1 and are listed as follows:

- Receiver Antenna: Used to receive the RF signal from the transmitter.
- Transmit/Receive (T/R) Device: Uses a T/R switch, circulator, diplexer, or duplexer to allow the same antenna for both the receiver and transmitter.
- Image Reject Filter: Reduces the image frequencies and frequencies in other bands from interfering with the incoming signal. It also reduces bandwidth and noise into the receiver.
- Low-Noise Amplifier (LNA): Main contributor to the noise figure (NF) of the receiver and provides gain.
- RF Downconverter: Converts RF signals to an IF using a LO for IF processing and filtering.
- Automatic Gain Control (AGC): Uses feedback techniques to automatically adjust the gain with respect to the incoming signal to keep a constant signal level into the rest of the receiver and detector. It also increases the dynamic range (DR) of the receiver.
- IF Downconverter: Converts the IF to a lower frequency or baseband for detection and signal processing.
- Analog-to-Digital (A/D) converter: Converts the incoming lower frequency or baseband to digital signals for digital signal processing (DSP).
- Digital Signal Processor: Provides data detection, despreading, and demodulation of the incoming digital waveform and retrieves the desired data information.

3.3 Receiver antenna

The receiver antenna gain is computed in the same way as for the transmitter with the calculation performed using the link budget in Chapter 1. Factors to consider in determining the type of antenna to use are frequency, the amount of gain required, and the size. These factors are similar to the ones used to determine the transmitter antenna. Parabolic dishes are frequently used at microwave frequencies. In many communications systems, the receiver uses the same antenna as the transmitter, which reduces the cost and size of the system. The antenna provides gain in the direction of the beam to reduce power requirements. It reduces the amount of received noise into the antenna which increases the sensitivity of the receiver. It also reduces the amount of interference and jamming into the receiver. The gain of the antenna is usually expressed in dBi, which is the gain in dB reference to what an ideal isotropic radiator antenna would produce.

3.4 Transmit/receive device

As was described in the transmitter section, the signal is received by the antenna and is passed through a device—a duplexer or diplexer, T/R switch, or circulator—to allow the same antenna to be used by both the transmitter and receiver. The chosen device must provide the necessary isolation between the transmitter circuitry and the receiver circuitry to prevent damage to the receiver during transmission.

3.5 Image reject filter

In nearly all receivers, an image reject filter/band reject filter is placed in the front end so that the image frequencies along with other unwanted signals are filtered out, with a decrease in noise. These unwanted signals are capable of producing intermodulation products (intermods). Intermods are caused by the nonlinearities of the LNA and the mixer.

The image frequencies are the other band of frequencies that, when mixed with the LO, will fall in the passband of the IF band. The image frequency has the same IF Frequency after the down conversion and it interferes with the desired signal at the IF frequency. This image frequency needs to be reduced before the mixer (Figure 3.2). For example,

LO = 1,000 MHz
Desired input signal = 950 MHz
IF = 1,000 − 950 MHz = 50 MHz

Thus,

Image frequency = 1,050 MHz
1,050 − 1,000 MHz = 50 MHz = same IF

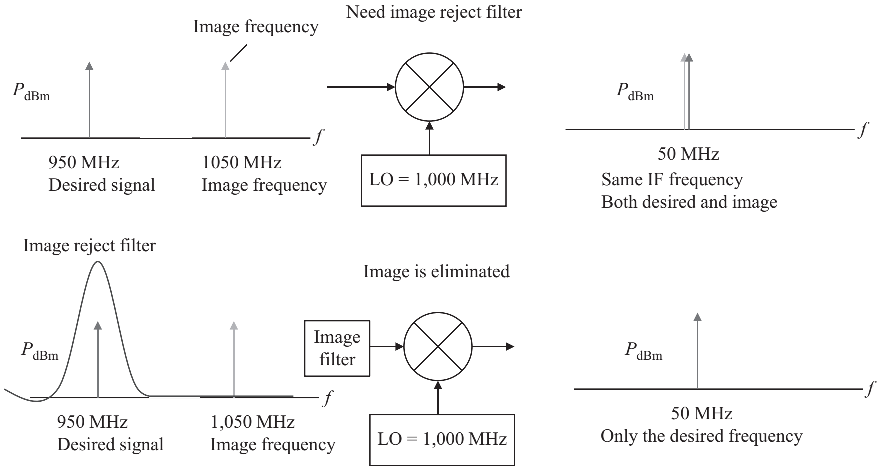


Figure 3.2 Image reject filter required to eliminate image frequency to prevent distortion

Therefore, if the receiver did not have an image reject filter in the front end, then an interference signal at 1,050 MHz would fall in the IF passband and jam the desired signal (Figure 3.2). In addition, the noise power in the image bandwidth would increase the receiver NF by 3 dB. Using an image reject filter or a band-pass filter (BPF) around the desire signal reduces the effects of the image frequency (Figure 3.2).

If the LO is changed to a different frequency, then the image frequency will change. For example,

LO = 900 MHz
Desired signal input = 950 MHz
IF = 950 − 900 MHz = 50 MHz

Thus,

Image frequency = 850 MHz
900 − 850 MHz = 50 MHz = the same IF

The image reject filter would be at 850 MHz (Figure 3.3).

Image reject mixers are sometimes used to reduce the image frequency without filtering. This is useful in a receiver system that contains a wide bandwidth on the RF front end and a narrow bandwidth IF after the mixer so that the image frequency cannot be filtered. This eliminates the need for double downconversion in some types of receivers. However, double downconversion rejects the image frequency much better. This is generally a trade-off on the image frequency rejection and cost. Also, the performance and the versatility of the super-heterodyne receiver are usually preferred.

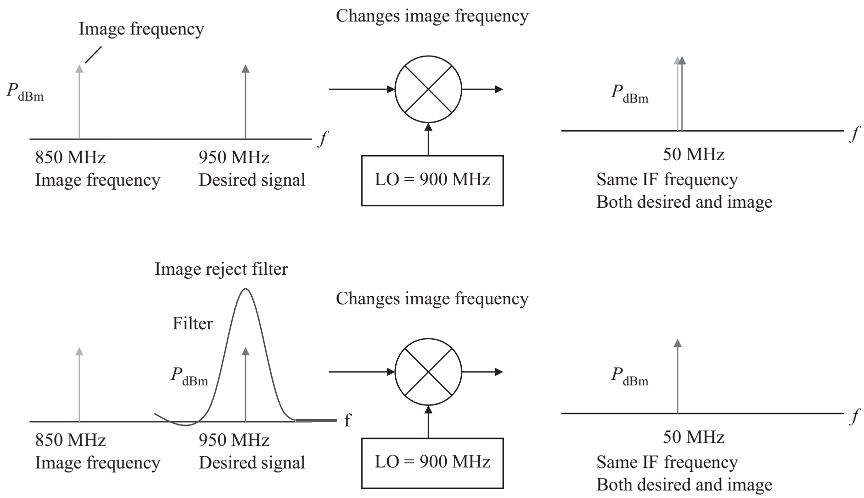


Figure 3.3 Changing LO changes the image frequency

3.6 Low-noise amplifier

The first amplifier in the receiver is called the Low-Noise Amplifier or LNA and is the major contributor to the NF of the receiver providing that the losses between gain elements are not too large and the bandwidth does not increase. An LNA is used to calculate the NF of the entire receiver using the Friis noise equation;

$$\text{Noise factor } F_s = F_1 + (F_2 - 1)/G_1 + (F_3 - 1)/G_1G_2 \ldots$$

Therefore, the NF would be:

$$NF = 10 \log F_s \approx 10 \log F_1, \text{if } G_1 \text{ is large}$$

For example, given the following parameters:
LNA amplifier:

$NF_1 = 6$ dB, $F_1 = 4$
$G_1 = 25$ dB, $G_1 = 316$

Mixer:

$L = 8$ dB (conversion loss), $L = 6.3$

IF amplifier:

$NF_2 = 6$ dB, $F_2 = 4$
$G_2 = 20$ dB, $G_2 = 100$

Therefore, the calculation for the NF of this receiver chain would be

$$F_t = F_1 + [(F_2 \times \text{Losses}) - 1]/G_1 = 4 + (4 \times 6.3 - 1)/316 = 4.08$$
$$\text{NF} = 10\log(4.06) = 6.10\,\text{dB}$$

An LNA also provides isolation between the LO used for downconversion and the antenna. This prevents LO bleed-through from appearing at the antenna port. This can be a problem for two reasons. If the system is designed for a low probability of intercept, then this LO signal is being transmitted from the antenna and can be detected. The other reason is this signal might interfere with other users.

Another benefit of using an LNA in the receiver is that less power is required from the transmitter if the NF is low. For a given desired signal-to-noise ratio (SNR) or E_b/N_o, the NF alters these according to the link budget, as specified in Chapter 1. In general, an LNA amplifies the desired signal with minimum added noise and establishes the receiver noise level. However, there are some reasons for not using an LNA for special applications as follows:

- Low-noise amplifiers reduce the DR of a receiver, which is the operating range from the minimal signal to the maximum signal that can be detected and processed. For example, if the 1 dB compression point is +10 dBm, the maximum input power to an LNA with a gain of 25 dB is −15 dBm. Without a preamp, the 1 dB compression for a low-level mixer is +3 dBm. Also, the LO drive can be increased to provide an even higher DR without the LNA.
- The interference signals, clutter, and bleed-through from the transmitter may be more significant than the noise floor level, and therefore establishing a lower NF with an LNA may not be required.
- Cost, space, DC power consumption, and weight are less without a preamplifier.
- Some airborne radar systems use mixer front ends (no preamp or LNA) to reduce weight.

The results show that an LNA gives a better NF, which results in better sensitivity at the cost of DR, cost, and space. An LNA also provides isolation between the LO and the antenna. Also, the low NF established by the LNA decreases the power requirement for the transmitter. In general, an LNA in the receiver provides the best overall design and should be included and carefully designed for the best NF.

3.7 RF downconverter

After the LNA, the received RF signal is downconverted to a lower frequency or IF for processing. This is accomplished by using an LO, a mixer, and a BPF as shown in Figure 3.1. The IF BPF is used to reduce the harmonics, spurious, and intermod products that are produced in the downconverter process. The IF is usually a lower frequency than the incoming frequency; however, in some applications the IF could be higher, which would require an RF upconverter. Generally, most receivers use an RF downconverter.

3.8 Mixers

The downconversion process requires a mixer to translate the frequency carrier to an IF and eventually to a baseband frequency. Many types of mixers and their characteristics should be examined when developing an optimal downconversion system.

3.8.1 Mixer spur analysis—level of spurious responses and intermods

A mixer spur analysis is done for each mixer output to

- Determine which mixer products fall in the passband of the output;
- Determine the frequencies to be used to ensure that mixer spurs do not fall in the passband.

Spurious signals are generated when mixing signals up and down in frequency. They are the mixer products and are designated as $n \times m$-order spurs, where n is the harmonics of the LO and m is the harmonics of the RF or IF (the input of the mixer could be an RF signal or an IF signal depending on the frequency translation). These mixer spurs can cause problems if they fall in the passband since they cannot be filtered.

A spurious analysis should be done to determine where the spurs are located—if they fall in the passband—and the power of the spur with respect to the desired signal. Several software programs have been written to assist in determining where the spurs are for a given system. These programs take two input frequencies and multiply them together, depending on the order specified, to generate the possibilities. For example, Table 3.1 shows a sixth-order system analysis where both the sum and difference need to be evaluated:

The first number multiplies the LO and the second number multiplies the input. The resultants are added and subtracted to determine the frequency of the spurs. Note that the 1×1 contains the desired signal and a spur, depending whether or not the wanted signal is the sum or the difference. Usually, this is obvious and the unwanted signal is filtered out.

Selecting the right LO or specifying the operational frequencies for a selected level of spur rejection is done by comparing all the possible spur locations with respect to a given amplitude threshold. This is dependent on high-side, low-side, sum, or difference. Frequency selection depends on mixer spur analysis.

3.8.2 Sixth-order analysis

A sixth-order analysis is standard for spur analysis. This ensures that the spurious signals are about 60 dB down from the desired output for most mixers. The analysis below is for a sixth-order spur analysis with the desired signal being the difference frequency. The LO is higher in frequency than the highest desired frequency, which eliminates the spurs 2×1, 3×1, 4×1, 5×1, 3×2, and 4×2.

Table 3.1 Sixth-order analysis; both the sum and difference need to be evaluated

$1 \times 0 = \text{LO}$	$0 \times 1 = \text{Input}$
Carrier (sometimes in reverse)	
$2 \times 0 =$	$0 \times 2 =$
$3 \times 0 =$	$0 \times 3 =$
$4 \times 0 =$	$0 \times 4 =$
$5 \times 0 =$	$0 \times 5 =$
$6 \times 0 =$	$0 \times 6 =$
$1 \times 1 = \text{Desired}$	
$2 \times 1 =$	$1 \times 2 =$
$3 \times 1 =$	$1 \times 3 =$
$4 \times 1 =$	$1 \times 4 =$
$5 \times 1 =$	$1 \times 5 =$
$3 \times 2 =$	$2 \times 3 =$
$4 \times 2 =$	$2 \times 4 =$
$2 \times 2 =$	$3 \times 3 =$

The worst case highest in-band frequency for the 1×0 spur would be $\text{LO} - f_\text{l}$. Since it is high-side injection, the 1×0 (LO) would always be greater than $\text{LO} - f_\text{l}$:

$$1 \times 0;\ \text{LO} > \text{LO} - f_l:$$

This also applies for the following spurs: 2×0, 3×0, 4×0, 5×0, 6×0.

The worst case lowest in-band frequency for the 1×0 spur would be $\text{LO} - f_\text{h}$. Since it is high-side injection, the 1×0 (LO) would always be less than $\text{LO} - f_\text{h}$:

$$1 \times 0; \text{LO} < \text{LO} - f_\text{h}:$$

This also applies for the following spurs: 2×0, 3×0, 4×0, 5×0, 6×0.

Note, the bandwidth is from $\text{LO} - f_\text{h}$ to $\text{LO} - f_\text{l}$, which none of these spurs fall into.

The analysis continues for all possible spur products to see if they fall into the operating band. If they do, then the spur chart is examined to determine how big the spurious signal will be. If the design is not to allow sixth-order products in the band, then each of the possible products must be calculated to ensure that none of them fall into the desired bandwidth. The same type of mixer analysis can be done for each mixer configuration to ensure that the spurious signals do not fall in the desired band and can be filtered out of the system.

Another point to make about mixers is that, in general, the voltage standing wave ratio (VSWR) for a mixer is not very good, about 2:1. If the source is 2:1, then, for electrical separation of greater than one-fourth wavelength, the VSWR can be as much as 4:1. This is equivalent to having a mismatched load four times larger (or one-fourth as large) than the nominal impedance.

3.8.3 High-level or low-level mixers

High-level mixers use higher voltages and require more power to operate than do the standard low-level mixers. Typical values are +13, +17, and +28 dBm. The advantages of using high-level mixers include the following:

- High third-order intercept point
- Less cross modulation
- More suppression of two-tone third-order response, $(2f_1 - f_2) \pm$ LO.
- Larger DR
- Lower conversion loss, better NF
- Best suppression above bottom two rows of a mixer spurious chart (Table 3.2).

Low-level mixers use lower voltages and require less power to operate than do the standard high-level mixers. Typical values are +0, and +7 dBm. The advantages of using low-level mixers include the following:

- Less complex, less expensive, less system power
- Easier to balance—better isolation between ports
- Lower DC offset
- Less mixer-induced phase shift.
- More covert, less LO bleed-through.
- Best suppression of bottom two rows of mixer chart (Table 3.3).

3.8.4 High-side or low-side injection

Another consideration when selecting a mixer is whether to use high-side vs low-side injection. High side refers to the LO being at a higher frequency than the input signal frequency; low side refers to the LO being at a lower frequency than the input signal frequency. Some of the main points to consider in the selection are as follows:

- Performing a spurious signal analysis assists in determining the best method to use.
- High-side injection inverts the signal; lower sideband becomes upper sideband, which may cause some problems.

Table 3.2 Typical spurious chart for a high-level mixer

Harmonics of f_R									
	3	>90	>90	>90	>90	>90	>90	>90	89
	2	73	83	75	79	80	80	77	82
	1	24	0	34	11	42	18	49	37
	0		18	10	23	14	19	17	21
	Harmonics of f_L	0	1	2	3	4	5	6	7

Table 3.3 Typical spurious chart for a low-level mixer

Harmonics of f_R								
5	80	79	78	72	82	71	x	x
4	88	80	90	80	90	82	x	x
3	58	55	65	54	66	54	x	x
2	69	64	71	64	73	61	x	x
1	16	0	25	12	33	19	x	x
0		24	30	37	41	34	x	x
Harmonics of f_L	0	1	2	3	4	5	6	7

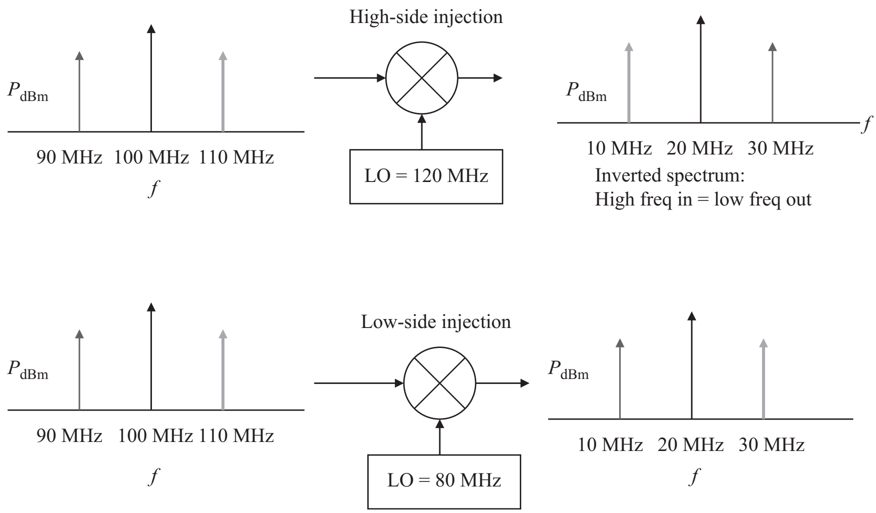


Figure 3.4 High-side vs low-side injection reverses bands and changes image frequency

- Low-side injection provides a lower frequency for the LO, which may be easier to get and less expensive.
- The image frequency is different and analysis needs to be done to determine which of the image frequencies will affect the receiver the most (Figure 3.4).

Another possibility to prevent inversion of the waveform is to do a double high side conversion (Figure 3.5). Since this inverts the signal twice, the net result is a noninverted spectrum.

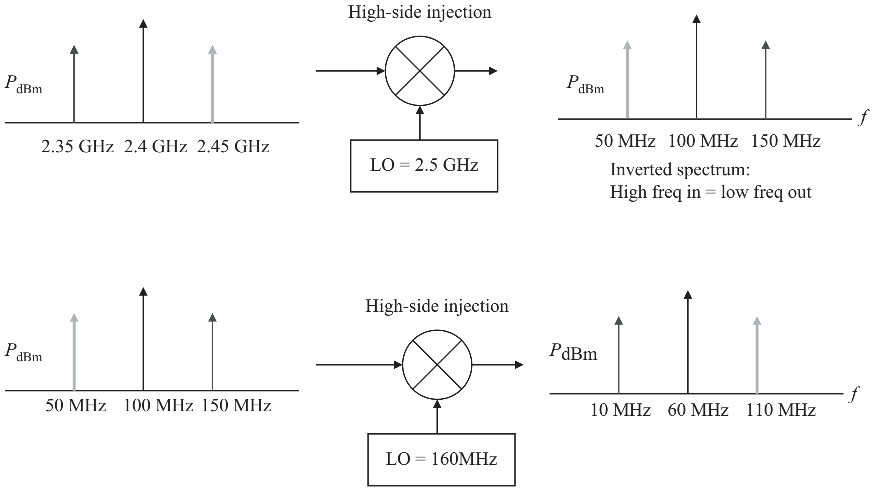


Figure 3.5 Double high-side conversion prevents inversion

3.9 Automatic gain control

Automatic gain control is shown in the IF section of the receiver (Figure 3.1). However, the AGC can be placed in the RF, IF, baseband, digital processor, or multiple places in the receiver where it is important to keep the output levels constant, to increase the DR, and to automatically adjust the gain with respect to the incoming signal. The AGC uses feedback to detect the level of the incoming signal, and automatically controls the receiver gain to the desired level of the output signal. This keeps the signal level constant for detection and signal processing and increases the range of signals that the receiver can handle. A detailed discussion on AGC can be found in Chapter 4.

3.10 IF downconverter

The IF downconverter is used to convert the IF to a lower frequency or baseband. This reduces the sample time of the analog-to-digital converter (ADC) and digital processing and is composed of an LO, mixer, and low-pass filter (LPF), the latter of which limits the high-frequency components to allow the ADC to sample and process the signal. This second downconverter creates a superheterodyne (double conversion) receiver, which is ideal to prevent and filter out unwanted harmonics, spurious, and intermod products to eliminate interference in the digital signal processer.

The double downconversion process relaxes constraints on filters and allows for common circuitry for different systems. For example, a receiver operating at 1 GHz and another operating at 2 GHz could both downconvert the signal to a common IF of 50 MHz, and the rest of the receiver could be identical.

Some of the newer receivers digitize the IF band directly and do a digital baseband downconversion before processing. This is especially true for quadrature downconversion using an in-phase (I) channel and a quadrature-phase (Q) channel. This relaxes the constraints of quadrature balance between the quadrature channels in the IF/analog portion of the receiver, since it is easier to maintain quadrature balance in the digital portion of the receiver and reduces hardware and cost.

Quadrature downconversion is a technique that is used to eliminate unwanted bands without filtering. For example, during a conversion process, both the sum and difference terms are created as shown:

$$A\cos(\omega_s)t \times B\cos(\omega_c)t = AB/2\cos(\omega_s+\omega_c)t + AB/2\cos(\omega_s-\omega_c)t$$

The receiver processes the incoming signal using a quadrature downconverter so that when the worst case situation occurs with the sin(ω_c) multiplying, the quadrature channel will be multiplying with a cos(ω_c) so that the signal is recovered in either (or both) the I and Q channels as shown:

$$\begin{aligned}\text{I channel} &= [AB/2\cos(\omega_s+\omega_c)t + AB/2\cos(\omega_s-\omega_c)t][\sin(\omega_c)t]\\ &= AB/4[\sin(\omega_s+2\omega_c)t + \sin(\omega_s)t - \sin(\omega_s)t + \sin(\omega_s-2\omega_c)t]\\ &= AB/4[\sin(\omega_s+2\omega_c)t + \sin(\omega_s-2\omega_c)t]\end{aligned}$$

$$\begin{aligned}\text{Q channel} &= [AB/2\cos(\omega_s+\omega_c)t + AB/2\cos(\omega_s-\omega_c)t][\cos(\omega_c)t]\\ &= AB/4[\cos(\omega_s)t + \cos(\omega_s+2\omega_c)t + \cos(\omega_s-2\omega_c)t\cos(\omega_s)t]\\ &= AB/4[\cos(\omega_s+2\omega_c)t + \cos(\omega_s)t + \cos(\omega_s)t + \cos(\omega_s-2\omega_c)t]\\ &= AB/2\cos(\omega_s)t + AB/4\cos(\omega_s+2\omega_c)t + AB/4\cos(\omega_s-2\omega_c)t\end{aligned}$$

This shows all of the signal being received in the Q channel, cos(ω_s), and none in the I channel. Most of the time this will be a split, and the magnitude and phase will be determined when combining the I and Q channels. This quadrature technique uses carrier recovery loops when receiving a suppressed carrier waveform. For example, Costas loops, which are commonly used in direct sequence systems, use this technique to recover the carrier and downconvert the waveform into I and Q data streams.

3.11 Splitting signals into multiple bands for processing

Some receivers split the incoming signal into multiple bands to aid in processing. This is common with intercept-type receivers that are trying to detect and intercept an unknown signal. If the signal is divided up in multiple bands for processing, then the signal is split up but the noise remains the same if the bandwidth has not changed. Therefore, the SNR will degrade each time the signal is split by at least 3 dB. However, if the signal is coherent and is summed coherently after the split, then there is no change in the SNR except some loss in the process gain. Thus, careful examination of the required SNR and the type of processing used with this type of receiver will aid in the design.

3.12 Bandwidth considerations

The bandwidth of the transmitted and received signal can influence the choice of IF bands used. The IF bandwidth is selected to prevent signal foldover. In addition, the IF bandwidth needs to reduce the higher frequencies to prevent aliasing. Also, the bandwidths need to be selected to relax design constraints for the filters.

For example, a direct sequence spread bandwidth is approximately 200 MHz wide. Since the total bandwidth is 200 MHz, the center RF needs to be greater than twice the 200 MHz bandwidth so that the unwanted sidebands can be filtered. An RF of 600 MHz is chosen because of the availability of parts. This also provides sufficient margin for the filter design constraints. If the bandwidth is chosen so that the design constraints on the filters requires an unrealistic shape factor (SF), or the frequency response of the filter is too great, then the filters may not be feasible to build. Bandwidth plays an important function in the SNR of a system. The narrower the bandwidth, the lower the noise floor, since the noise floor of a receiver is related to $kTBF$, where B is the bandwidth.

3.13 Phase noise

Phase noise is generally associated with oscillators or phase-locked loops (PLLs). Phase noise is the random change in the phase or frequency of the desired frequency or phase of operation. The noise sources include thermal noise, shot noise, and flicker noise or $1/f$ noise.

The different types of phase and frequency noise and the causes that generate each of the processes and definitions are shown in Figure 3.6. The noise beyond about 10 kHz is dominated by white noise, which is generally referred to as kTB noise since it is derived using Boltzmann's constant, temperature, and bandwidth. The $1/f$ noise or flicker noise is the dominant noise source from about 100 Hz to 10 kHz. This noise is produced by noisy electrical parts, such as transistors and amplifiers, and choosing a good LNA can help. The next type of noise, which is the

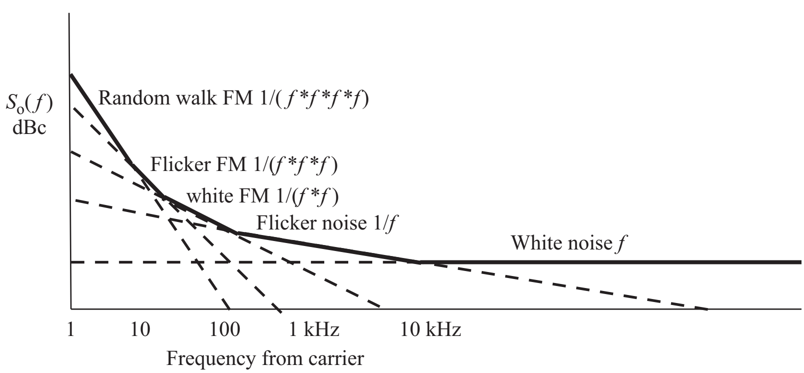


Figure 3.6 Phase noise analysis

dominant noise from about 20 to 100 Hz, is white FM noise, caused by passive resonator devices such as cesium and rubidium. White FM noise appears to change the frequency with respect to white noise and is designated as 1/*ff*. The next type of noise is the physical resonance mechanism or the actual parts in oscillators and is close to the carrier, about 5 to 20 Hz. This type of noise is known as flicker FM and is designated as 1/*fff*. The closest noise to the actual carrier frequency is 1/*ffff* noise, which is caused by vibration, shock, temperature, and other environmental parameters. This is called random-walk FM and is very difficult to measure since it is so close to the carrier and requires very fine resolution.

3.14 Filter characteristics

Filters are used to reject spurious signals, inter-modulation products, harmonics, jammers and cosite interference, adjacent channel users, image frequencies, and all other unwanted signals. They also are used to limit the noise bandwidth, set the detection bandwidth to prevent aliasing, and all out-of-band transmissions.

The filter roll-offs are often characterized by their SFs (Figure 3.7). The SF is defined as:

$$\mathrm{SF} = f(-60\ \mathrm{dB})/f(-3\ \mathrm{dB})$$

where $f(-60$ dB) is the frequency that is 60 dB down from the center of the passband and $f(-3$ dB) is the frequency that is 3 dB down from the center of the passband.

For example,

$$\mathrm{SF} = 20/10\ \mathrm{MHz} = 2\mathrm{MHz}$$

where $f(-60$ dB) is the 20 MHz and $f(-3$ dB) is the 10 MHz.

Note that these values may be evaluated at different levels for different vendors.

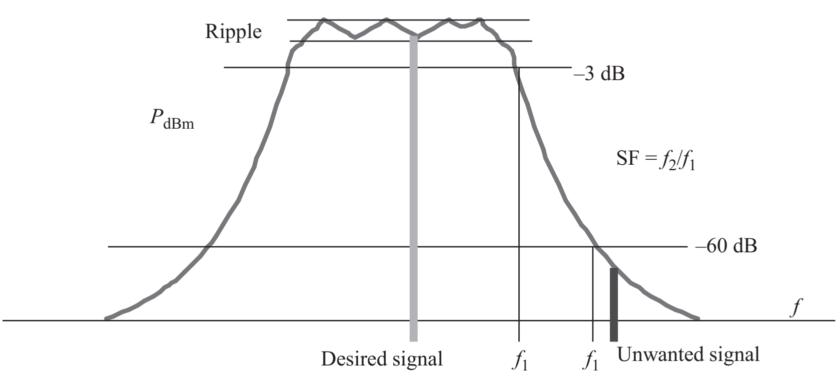


Figure 3.7 Typical filter response showing insertion loss and shape factor

Typical achievable SFs range from 2 to 12, although larger and smaller SFs can be achieved. SFs as low as 1.05 can be achieved with crystal filters. Note that insertion loss generally increases as the SF decreases. Insertion loss ranges from 1.2 to 15 dB. Surface-acoustic-wave filters achieve even a better SF at the expense of greater insertion loss and cost. They are also more prone to variations from vibration.

The percent bandwidth is used to specify the limitations of the ability to build a filter. The percent bandwidth is a percentage of the carrier frequency. For example, 1% bandwidth for 100 MHz would be 1 MHz bandwidth. This is usually specified as the 3 dB bandwidth.

If the bandwidth is not specified, then the 3 dB bandwidth is used. Typical achievable bandwidths using crystal filters for 10 kHz to 200 MHz range from 0.0001% to 3%.

Filters are used primarily for out-of-band rejection. This helps reduce interference and adjacent channel interference for multiple users. Many different types of filters can be used depending on the type of roll-off, phase distortion, passband characteristics, and out-of-band sidelobes. Some filters may have good roll-off characteristics but may experience sidelobes in the out-of-band response.

Caution needs to be used for filters that may act differently at higher frequencies, where the capacitors may look inductive and the inductors may look capacitive. This is usually specified at the resonance frequency of the passive component. The self-resonance frequency (SRF) of a component is where the impedance of the capacitance and the impedance of the inductance cancel out to provide low resistive impedance. The SRF should be much higher than the frequency at which the part is going to be used.

Another consideration in selecting a filter is the amount of amplitude variation in the passband, known as passband ripple. For example, Butterworth filters have low passband ripple, whereas Chebyshev filters specify the passband ripple that is desired.

Insertion loss is another key design consideration. Usually, the higher order filters and steeper roll-offs have higher insertion losses that have to be accounted for in the design of the receiver. A typical filter with a specified insertion loss and SF is shown in Figure 3.7.

3.15 Group delay

Group delay is the measurement of the delay of the frequency components through a device or system. This is an important consideration for digital communications. Digital signals and waveforms are equal to a sum of multiple frequencies with different amplitudes that are specified using the Fourier series. As these frequencies propagate through the device or system, if all of the frequency components have the same delay, then they are said to have constant group delay and the output digital signal will be the same as the input digital signal. However, if these frequencies have different delays through the device or system, which is nonconstant group

delay, then they will not add up correctly on the output and the output digital signal will be distorted (Figure 3.8). This distortion is known as dispersion and will cause intersymbol interference (ISI). This is where one symbol or digital signal interferes with another symbol or digital signal because they are spread out in time due to dispersion. Devices in a receiver such as filters do not have constant group delay, especially at the band edges (Figure 3.9). Some filters are better than others as far as group delay; for example, Bessel filters are known for their constant group delay. Constant group delay and linear phase are related since the group delay is the

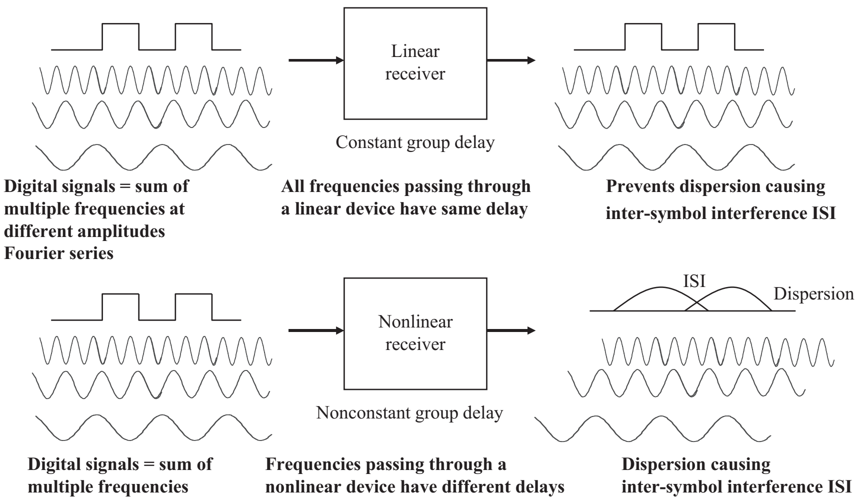


Figure 3.8 Group delay for digital signals

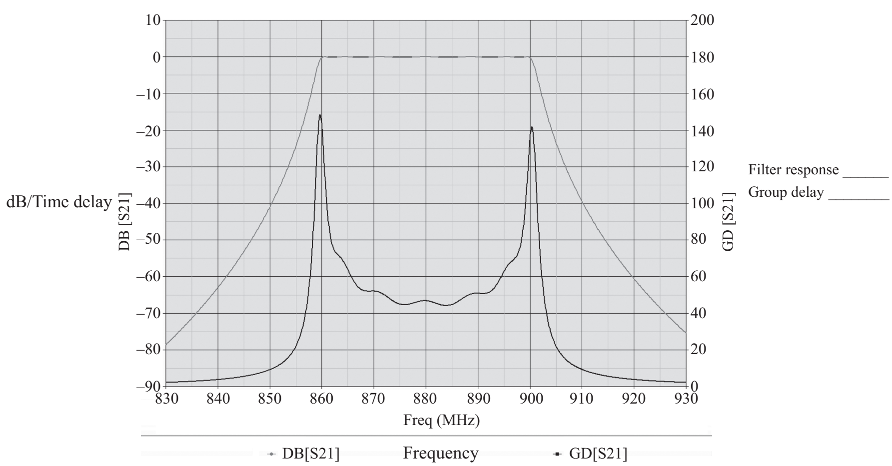
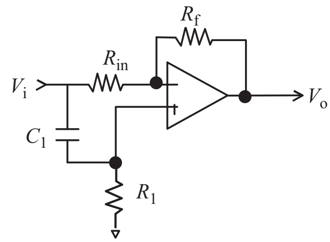
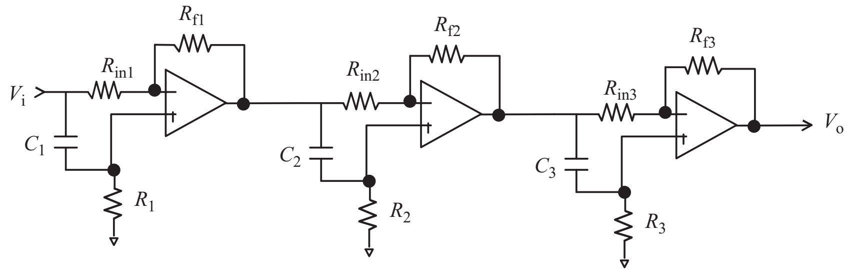


Figure 3.9 Group delay through a filter



Basic all pass network



Cascaded all pass network

Figure 3.10 Op amp all pass group delay equalizer

derivative of the phase. Thus, for linear phase, which means a constant slope, the group delay or the derivative of the constant slope is a constant number.

Some systems utilize group delay compensation to help mitigate the non-constant group delay of a device or system. The basic technique measures the group delay/phase linearity of the device or system. Several components affect group delay, including filters, amplifiers, and narrowband devices. Once the measured group delay has been obtained, an all-pass phase linearizer/group delay compensator can be used to compensate for the delay differences (Figure 3.10).

These compensators have no effect on magnitude, only phase, with their poles and zeros spaced equally and opposite of the $j\omega$ axis. The next step is to convolve the curves of the measured group delay and group delay compensator to produce a constant group delay for the system (Figure 3.11).

3.16 Analog-to-digital converter

An ADC is a device that samples the analog signal and converts it to a digital signal for digital processing. It must sample at least as fast as the Nyquist rate, or twice as fast as the highest frequency component of the signal. One of the problems that occur when an analog signal is digitized is the low resolution of the step size. The error associated with the step is called the quantization error.

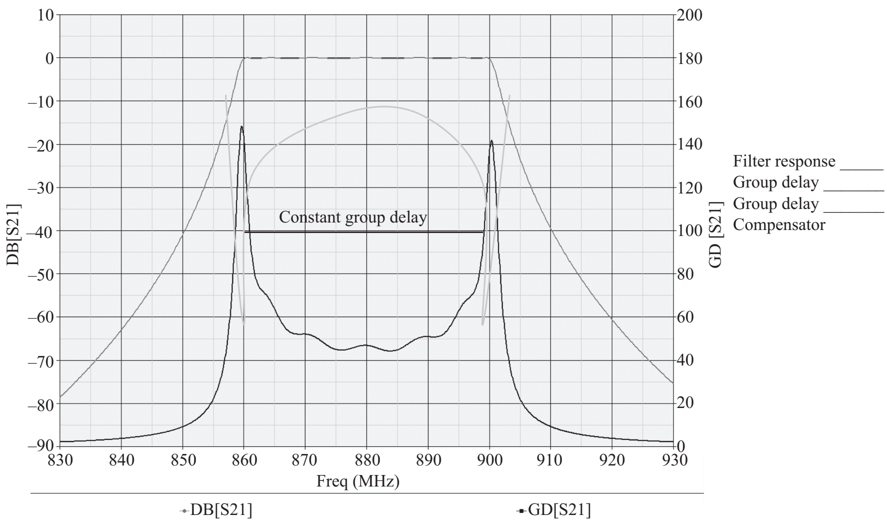
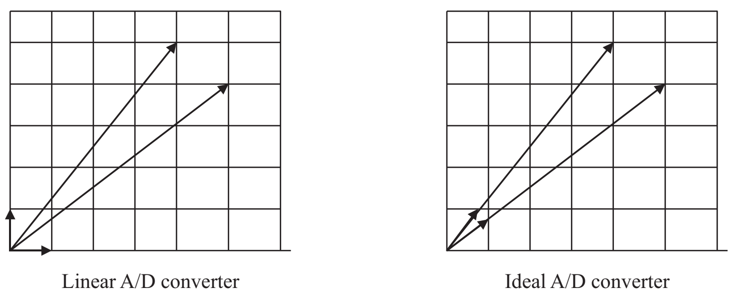


Figure 3.11 Group delay compensator resulting in a constant group delay



Ideal A/D converter percent error is reduced for lower level signals
μ-law and linear approximation designed to approximate ideal A/D converter

Figure 3.12 Quantization error for large and small signals using a linear vs an ideal A/D converter

As the signal becomes smaller in a linear ADC, the error becomes larger. Note that if the signal is large, the quantization errors in both I and Q channels produce a small percentage change in amplitude and phase. The phase error is just a few degrees, and the amplitude is a very small percentage of the actual amplitude, as shown in Figure 3.12. As for the smaller signal, quantization error in the I and Q channels produces a large percentage error in phase.

Most ADCs are linear, which makes it harder to detect smaller signals. To improve the sensitivity of the ADC, a log amplifier is sometimes used before the ADC or an ADC with a log scale (μ-law) is used.

This gives a higher response (gain) to lower level signals and a lower response (gain) to higher level signals and approximates the ideal ADC (Figure 3.12). Another way to improve the sensitivity of lower level signals is to use a piecewise linear solution where there are multiple ADC stages with different scaling factors. Each finer resolution stage is set to cover the LSBs of the previous ADC. This method provides a large DR while still being a linear process. Also, a piecewise μ-law solution can be used to increase the DR of the ADC. Noise dithering or adding noise to the input to the ADC improves SFDR with only a small degradation in the SNR. This randomizes the LSB's of the ADC reducing or spreading out the spurs that are present in the ADC. Also, the noise can be inserted out-of-band to maintain the SNR. For each bit of the ADC, the DR is increased by approximately 6 dB/bit. The reason for this is that the DR of the voltage for each bit in the ADC is divided in half and then converted to dB:

$$10\log(1/2)^2 = -6\text{ dB}$$

For each ADC bit, the value of the bit could be high or low, which splits the decision by 1/2. For example, if there is a range from 0 to 10 V on the input to the ADC and there is 1 bit in the ADC with a threshold set at 5 V, then the DR has been increased by 1/2. Now the ADC can detect 5 and 10 V instead of just 10 V. If the bit value is high, then the range is from 5 to 10 V; if the bit value is low, then the range is from 0 to 5 V. Since voltage levels are being evaluated, the 1/2 is squared. If there are two bits, then the range can be divided into four levels, and each level is one-fourth the total DR, which calculates to −12 dB.

The example is summarized as follows:

- Voltage range reduced by 1/2
- 20 log(1/2) = −6 dB
- Example: 0 to 10 V detection range
  - 1 bit reduces voltage range by 1/2
  - High "1" = 5 to 10 V
  - Low "0" = 0 to 5 V
  - Another bit reduces by 1/2
  - 0,0 = 0 to 2.5 V
  - 0,1 = 2.5–5 V
  - 1,0 = 5 to 7.5 V
  - 11 = 7.5 to 10 V
- 10 to 5 V = 6 dB
- 10 to 2.5 V = 12 dB

3.17 Sampling theorem and aliasing

Nyquist states that an analog function needs to be sampled at a rate of at least two times the highest frequency component in the analog function to recover the signal. If the analog function is sampled less than this, aliasing will occur (Figure 3.13).

When a signal is sampled, harmonics of the desired signal are produced. These harmonics produce replicas of the desired signal (positive and negative frequencies) in

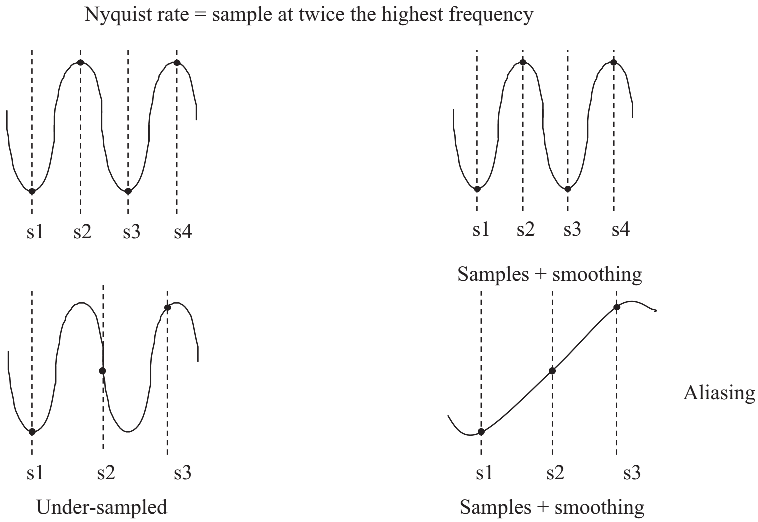


Figure 3.13 Sampling theorem—Nyquist theory and aliasing

the frequency domain at $1/T_S$ intervals. When the analog signals are undersampled, the harmonics of the repeated signal start to move into the passband of the desired signal and produce aliasing. In addition, any higher frequency component that is present at the input to the sampler, if it is undersampled, these components can also move into the passband and interfere with the desired signal due to aliasing. To illustrate the concept, if the highest frequency of the desired signal is 1 MHz, then the sample rate required is 2 million samples per second (Msps). If this frequency is sampled at this rate, the output of the sampler would be alternating plus and minus values at a 2-Msps rate (Figure 3.14). The reconstruction of the samples would produce a 1 MHz frequency. If there is an incoming signal at a higher frequency and it is sampled at the same rate (undersampled according to the Nyquist rate), then the output would produce a frequency that would appear as a lower frequency than the actual input frequency (Figure 3.15). This reconstructed frequency could fall into the desired signal bandwidth and interfere with the correct signal. Antialiasing or prealiasing filters can eliminate this problem for digital sampling and transmissions. This antialiasing filter is required before any sampling function to prevent aliasing of the higher frequencies back into the desired signal bandwidth. The high frequencies above the Nyquist criteria are filtered before they are sampled to prevent aliasing and distortion of the desired signal.

In some systems, this Nyquist under sampled technique is controlled and used to process desired signals at higher frequencies by undersampling. This allows for the sample rate of the detector to be slower and provides a means of processing much higher frequency signals with slower clocks. This can reduce cost and increase bandwidths for the detection process. For example, if the desired signal is the high-frequency signal in Figure 3.15, then the resultant aliased low-frequency signal would be the desired signal and is detected by this lower sample rate detector. However, careful design is needed so that unwanted signals do not alias and cause interference with the desired signal.

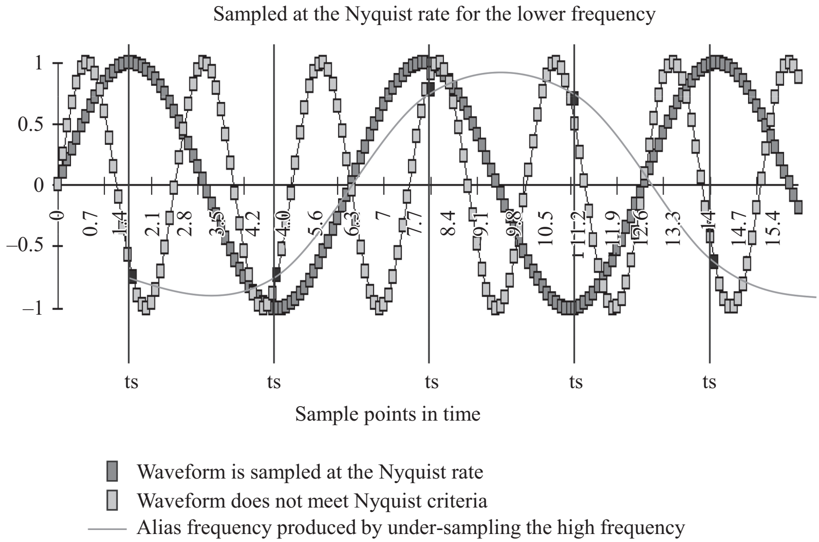


Figure 3.14 Graph showing the Nyquist criteria for sampling and aliasing

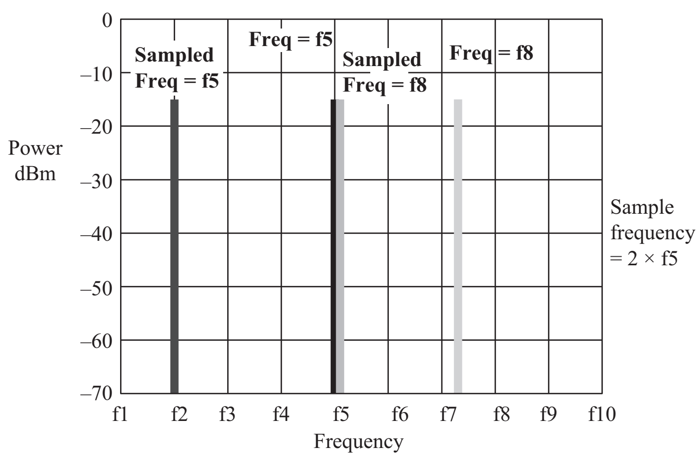


Figure 3.15 Frequency spectrum showing the Nyquist criteria for sampling and aliasing

3.18 Dynamic range/minimum detectable signal

There is a great deal of confusion in the definitions and measurements of MDS and DR. Care should be taken to ensure accurate analysis and to compare different systems with the same criteria. One of the main problems is that many systems

today process signals in DSP integrated circuits, which makes comparisons with the analog-type systems difficult. MDS is a measurement of how well a receiver can detect a signal with respect to noise and is generally calculated at the ADC. DSP systems may not be able to evaluate the signal at the ADC, especially if spread spectrum systems are used and despreading is performed in the digital domain. Before determining the MDS for a particular system, careful analysis needs to be performed to provide the optimal place in the system to do the calculation. Also, the criteria for calculating the MDS need to be considered to evaluate each system's architecture. BER, tangential sensitivity (TSS), where the pulse is tangential to the noise, SNR, and others can be used as the criteria. To evaluate each system fairly, the same process should be used for comparison.

DR can be calculated using the minimum signal that the ADC can detect and the maximum voltage that the ADC can process. This assumes there is no saturation and the ADC can handle the full DR. For example, if the maximum voltage is 2 V, then the maximum 2 V will fill up the ADC with all ''1''s. For each bit of the ADC, the DR is increased by approximately 6 dB/bit.

Consider, for example, a system with an ADC that can handle a maximum of 2 V and uses an 8-bit ADC. The 8-bit ADC provides 48 dB of DR which produces a minimum signal that is approximately −48 dB below 2 V. Therefore, the minimum signal in volts would be approximately 7.8 mV, with a DR equal to 48 dB.

The system noise, which establishes the noise floor in the receiver, is composed of thermal noise (*kTBF*) and source noise. The values of k, T, B, and F are usually converted to dB and summed together. Additional noise is added to the system due to the source noise, including LO phase noise, LO bleed-through, and reflections due to impedance mismatch. Consequently, the noise floor is determined by

$$\text{Noise floor} = \text{thermal noise} + \text{source noise}$$

The noise floor and the minimum signal at the ADC are used to calculate the SNR, and the MDS is calculated using given criteria for evaluation, such as the TSS for pulsed systems.

The input noise floor for the system is used to determine the amount of gain that a receiver must provide to optimize the detection process. The gain required from the receiver is the amount of gain to amplify this input noise to the threshold level of the ADC's least significant bit (LSB). Therefore, the gain required (G_r) is

$$\begin{aligned} G_r(\text{dB}) = {} & [\text{noise floor at the ADC's LSB(dB)}] \\ & - [\text{input noise floor at the receiver(dB)}] \end{aligned}$$

The DR of the receiver is the DR of the ADC unless additional DR devices are included, such as log amplifiers or AGC, to handle the higher level inputs. Many digital systems today rely on saturation to provide the necessary DR. The DR of the receiver depends on the portion of the receiver that has the smallest difference

between the noise floor and saturation. DR is another often misunderstood definition. The definition used here is the total range of signal level that can be processed through the receiver without saturation of any stage and within a set BER or defined MDS level. (However, some systems can tolerate saturation, which increases this dynamic.) This assumes that the signal does not change faster than the AGC loops in the system. If the signal changes instantaneously or very fast compared to the AGC response, then another definition is required, instantaneous DR (IDR). The IDR is the difference between saturation and detectability, given that the signal level can change instantaneously. The IDR is usually the DR of the ADC, since the AGC does not respond instantaneously.

To determine the receiver's DR, a look at every stage in the receiver is required. The NF does not need to be recalculated unless there is a very large amount of attenuation between stages or the bandwidth becomes wider with a good deal of gain. Generally, once the NF has been established, a look at the saturation point in reference to the noise level at each component can be accomplished. An even distribution of gains and losses is usually the best. A DR enhancer such as an AGC or a log amplifier can increase the receiver's DR but may not increase the IDR. However, devices such as log amplifiers can actually reduce the overall DR by adding noise to the desired signal.

Note that the DR can be reduced by an out-of-band large signal, which causes compression and can also mix in noise. The jamming signals should be tested and analyzed to determine the degradation of the receiver's DR.

3.19 Types of DR

DR in a receiver design can designate either amplitude or frequency. The two DRs are generally related and frequently only one DR is used, but the type of DR specified is given for a particular reason, for example, a two-tone test or receiver test.

3.19.1 Amplitude DR

If a design requires the output be within a given amplitude signal level, then the DR is generally given as the difference between the MDS and the maximum signal the receiver can handle (saturation). Often the 1 dB compression point provides a means of determining the maximum signal for the receiver. This point is where the output signal power is 1 dB less than what the linear output power is expected to be with a given input, which means there is some saturation in the receiver (Figure 3.16). Operating at this point results in 1 dB less power than what was expected.

If the DR needs to be improved, then an AGC or a log amplifier can be added. The amplitude DR is probably the most commonly used DR term. Compression and saturation are important to prevent distortion of the incoming signal, which generally increases the BER.

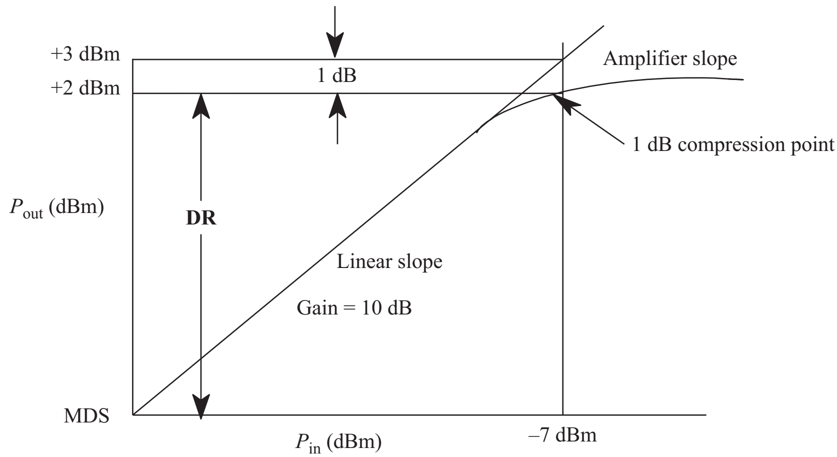


Figure 3.16 Amplitude DR using 1 dB compression point

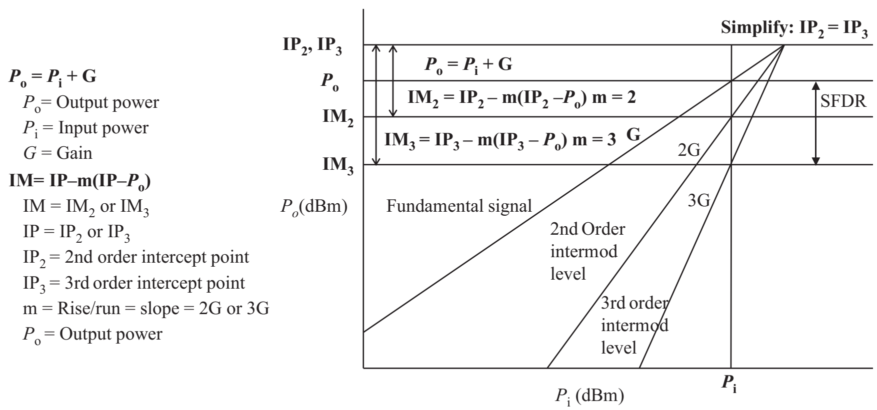


Figure 3.17 Frequency dynamic range—two-tone spur free dynamic range, SFDR

3.19.2 Frequency DR

Many systems are concerned with the frequency DR, which describes the ability of a receiver to distinguish different frequencies, whether it is between multiple desired frequencies or between the desired frequency and spurious signals or jammers. This is related to the 1 dB compression point DR, since when non-linearities are present there are unwanted spurious signals generated which reduce the frequency DR (Figure 3.17). Two methods of analysis typically prove useful information when considering frequency DR.

3.19.3 Single-tone frequency DR

One method is to use a one-tone DR and calculate the spurs that will reduce the frequency DR because of this tone. This method is not used very often because generally there is more than one input signal and the harmonics are out of the frequency band.

3.19.4 Two-tone frequency DR

A better and more common way is to define the DR using a two-tone DR analysis. This approach takes two input signals and their generated spurious products and calculates their power levels using intercept points on the linear or desired slope (Figure 3.17).

The first-order spurs—f_1 and f_2—do not test nonlinearity, so this method is not used. The second-order spurs—$2f_1$, $2f_2$, $f_1 \times f_2 = (f_1 + f_2, f_1 - f_2)$—are caused by nonlinearities and are at high power levels but can generally be easily filtered out. If they cannot be, since they are larger than third order, this could cause some problems due to the high power level of the spurs. The third-order spurs—$3f_1$, $3f_2$, $2f_1 \times f_2 = (2f_1 - f_2,\ 2f_1 + f_2)$, $f_1 \times 2f_2 = (f_1 - 2f_2,\ f_1 + 2f_2)$—are also caused by nonlinearities. These spurs are all filtered out except for $2f_1 - f_2, f_1 - 2f_2$, which are in band and are used to test the frequency DR.

The second-order intercept point helps to determine the second harmonics of the two inputs (2×0, 0×2) and the product of the two signals (1×1). The second-order spurs are the highest power intermod products that are created due to a nonlinear portion of the receiver. However, if the bandwidth is limited to less than an octave, these products are eliminated and the third-order spurs (2×1, 1×2) become the strongest products in-band. These spurs generally fall into the operating band, especially when they are close together in frequency.

The third-order intercept point becomes the criteria for calculating the DR for the receiver (Figure 3.17). Caution must be used when performing the analysis because if the bandwidth is greater than an octave then the second-order intermods cannot be neglected.

3.20 Second- and third-order intermodulation products

The second- and third-order intercept points are used to determine the intermod levels in the system. The output power (in dB) plotted against the input power is a linear function up to the compression point. The slope of this line is constant with gain; that is, $P_o(\text{dBm}) = P_i(\text{dBm}) + G(\text{dB})$. The slope of the second-order line is equal to twice that of the fundamental; that is, for every 1 dB of increase in output fundamental power, the second-order output signal power increases by 2 dB. The slope of the third-order line is equal to three times that of the fundamental; that is, for every 1 dB of increase in output fundamental power, the third-order output signal power increases by 3 dB. If the fundamental plot is extended linearly beyond the 1-dB compression point, then this line will intersect the second- and third-order lines. The point where the second-order curve crosses the fundamental is called the

second-order intercept point. The point where the third-order curve crosses the fundamental is called the third-order intercept point and is generally not at the same location as the second-order intercept point. The intercept points are given in dBm. The intermod levels can then be calculated for a given signal level and can be graphically shown and calculated as shown in Figure 3.17.

The second-order intermod signal will be two times farther down the power scale from the intercept point as the fundamental. Therefore, the actual power level (in dB) of the second-order intermod is

$$\mathrm{IM}_2 = \mathrm{IP}_2 - 2(\mathrm{IP}_2 - P_o)$$

where IM_2 is the second-order intermod power level (in dBm), IP_2 is the second-order intercept point, and P_o is the output power of the fundamental signal.

The third-order intermod signal will be three times farther down the power scale from the intercept point as the fundamental. Therefore, the actual power level (in dB) of the third-order intermod is

$$\mathrm{IM}_3 = \mathrm{IP}_3 - 3(\mathrm{IP}_3 - P_o)$$

where IM_3 is the third-order intermod power level (in dBm), IP_3 is the third-order intercept point, and P_o is the output power of the fundamental signal.

For example, suppose the second-order and third-order intercept points are at +30 dBm (most of the time they are not equal, however, for simplicity in the explanation assume $\mathrm{IP}_3 = \mathrm{IP}_2$). The input power level is at −80 dBm, and the gain is at +50 dB. The signal output level would be at −30 dBm. The difference between the intercept points and the signal level is 60 dB. The following example shows what the intermod signal levels are.

Given:

$\mathrm{IP}_2 = \mathrm{IP}_3 = +30$ dBm
$P_i = -80$ dBm
$G = +50$ dB

calculate:

$P_o = G \times P_i = 50\text{ dB} + -80\text{ dBm} = -30\text{ dBm}$
$\mathrm{IP}_2 - P_o = \mathrm{IP}_3 - P_o = +30\text{ dBm} - (-30\text{ dBm}) = 60\text{ dB}$
$P_o = \mathrm{IP}_{(2,3)} - 1(\mathrm{IP}_{(2,3)} - P_o) = +30\text{ dBm} - 1(60\text{ dB}) = -30\text{ dBm}$
$\mathrm{IM}_2 = \mathrm{IP}_2 - 2(\mathrm{IP}_2 - P_o) = +30\text{ dBm} - 2(60\text{ dB}) = -90\text{ dBm}$
$\mathrm{IM}_3 = \mathrm{IP}_3 - 3(\mathrm{IP}_3 - P_o) = +30\text{ dBm} - 3(60\text{ dB}) = -150\text{ dBm}$

which results in:

$\mathrm{IM}_2 = -90\text{ dBm} - (-30\text{ dBm}) = -60\text{ dBc}$ (60 dB down from P_o)
IM_2 is generally easy to filter out because second order products are generally not in the passband
$\mathrm{IM}_3 = -150\text{ dBm} - (-30\text{ dBm}) = -120\text{ dBc}$ (120 dB down from P_o) = Spur Free Dynamic Range (SFDR)

Since the second-order signals can be generally filtered out, the third-order products limit the system's DR.

IP_3 can be calculated by measuring P_o and IM_3 dBc as follows:

Measured on a spectrum analyzer: $P_o = -30$ dBm, $IM_3dBc = -120$ dBc
$IM_3 = IP_3 - 3(IP_3 - P_o) = -2IP_3 + 3P_o$
$IM_3dBc = IM_3 - P_o = (-2IP_3 + 3P_o) - P_o = -2IP_3 + 3P_o = -2(IP_3 - P_o)$
$IM_3dBc/-2 = IP_3 - P_o$ therefore; $IP_3 = P_o - IM_3dBc/2 = -30$ dBm – $(-120$ dBc$/2) = -30$ dBm $+ 60$ dBc $= 30$ dBm

So $IP_3 = 30$ dBm

3.21 Calculating two-tone frequency DR

A method of calculating DR is by comparing input power levels shown in Figure 3.18. This example is for third-order frequency DR. The input power that causes the output power (P_o) to be right at the noise floor is equal to P_1. This is the minimum input signal that can theoretically be detected. The input power (P_3) is the power that produces third-order intermod products at the same P_o as what is at the noise floor. An increase in input power will cause the third-order spurs to increase above the noise (Figure 3.18). (Note that the 3rd order spur level will increase in power at a rate of 3 times faster than the desired signal level as input power is increased.)

The following equations are simple slope equations:

$$(P_o - IP_3)/[P_1 - (IP_3 - G)] = 1$$
$$P_1 = P_o - G$$
$$(P_o - IP_3)/[P_3 - (IP_3 - G)] = 3$$
$$P_3 = (P_o + 2IP_3 - 3G)/3$$

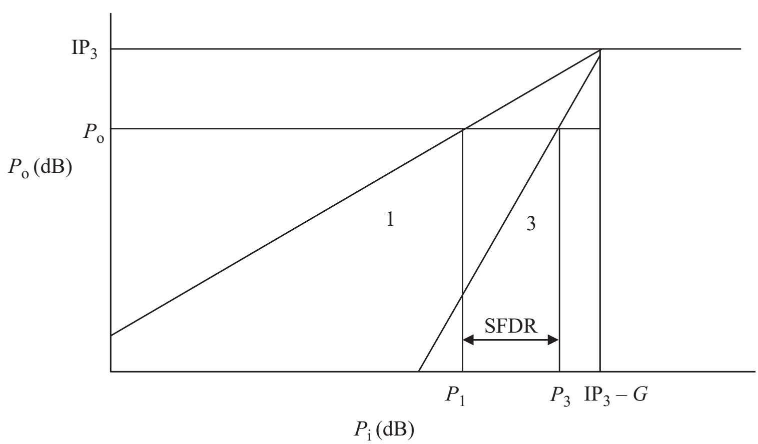


Figure 3.18 Frequency dynamic range analysis

The difference in the input powers gives the DR of the system:

$$\begin{aligned}\mathrm{DR} &= P_3 - P_1 = (P_o/3 + 2\mathrm{IP}_3/3 - G) - (P_o - G) = 2/3(\mathrm{IP}_3 - P_o)\\ P_o &= \text{noise floor} = -174 + 10\log B + \mathrm{NF} + G - G_p\\ \mathrm{DR} &= 2/3\left(\mathrm{IP}_3 + 174 - 10\log B - \mathrm{NF} - G + G_p\right)\end{aligned}$$

where DR is the dynamic range, IP_3 is the third-order intercept point, B is the bandwidth, NF is the noise figure, G is the gain, and G_p is the process gain.

Before the detector, the receiver NF bandwidth uses the smallest IF bandwidth for the calculation.

The previously given analysis gives the theoretical DR, but for the usable DR, a SNR needs to be specified. This requires the input power for the desired signal to increase sufficiently to produce the desired SNR. Since the desired spur level is still at the noise floor, this is a direct subtraction (in dB) of the desired SNR:

$$\mathrm{DR}_a = P_3 - P_{1a} = 2/3\left(\mathrm{IP}_3 + 174 - 10\log B - \mathrm{NF} - G + G_p\right) - \mathrm{SNR}$$

This is the usable third-order spurious-free dynamic range (SFDR).

From the same analysis, the second-order usable DR is:

$$\mathrm{DR}_a = 1/2\left(\mathrm{IP}_2 + 174 - 10\log B - \mathrm{NF} - G + G_p\right) - \mathrm{SNR}$$

To calculate the IP_3 for an amplifier, the two-tone test is used. The power output per tone is designated as P_{tone}. The level of the highest intermodulation (intermod) product is also specified or measured and is designated as intermodulation distortion (IMDs) in dBc (referenced to the carrier). Therefore, the equation to calculate IP_3 for both the input (IIP_3) and the output (OIP_3) is:

$\mathrm{OIP}_3 = P_{\text{out}} + \text{dBc}/2$: dBc is the highest level of the IMDs
$\mathrm{IIP}_3 = \mathrm{OIP}_3 - \text{gain (dB)}$

Here are two examples in calculating the OIP3 and the IMD level:

Example 1: Given: $P_{\text{tone}} = +5$ dBm, dBc = 30 dBc

Solve for OIP3:

$\mathrm{OIP}_3 = 5\text{ dBm} + 30\text{ dBc}/2 = 5\text{ dBm} + 15\text{ dB} = 20\text{ dBm}$

Example 2: Given: $\mathrm{OIP}_3 = 20$ dBm, $P_{\text{tone}} = +5$ dBm

Solve for IMD level:

$20\text{ dBm} = 5\text{ dBm} + X\text{ dBc}/2 = (20\text{ dBm} - 5\text{ dBm}) \times 2 = 30\text{ dBc}$

3.22 System DR

One concern in receiver system analysis is the determination of where the minimum DR occurs. To easily determine where this is, a block diagram of the receiver with the noise level and the saturation level is developed. This can be included as a parameter in the link budget. However, instead of cluttering the link budget, a separate analysis is usually performed. The outputs of the devices are listed with the noise level, the saturation level, and the calculated amplitude DR. This is shown in a spreadsheet in Table 3.4.

The procedure is to list all of the power levels and noise levels at every point in the system. The estimated NF is established at the LNA, which in turn estimates the relative noise floor. Note that the bandwidth of the devices is being used initially. If the bandwidth is narrowed down the line, then the noise is lowered by 10 log B. The top level is calculated by using either the 1-dB compression point or the third-order intercept point.

For the 1-dB compression point, the DR is calculated by taking the difference of the relative noise floor and this point. For the third-order intercept point method, the DR is calculated by taking two-third of the distance between the relative noise floor and the intercept point. The minimum distance is found and from this point the gains are added (or subtracted) to determine the upper level input to the system. The difference between the input relative noise and the upper level input is the DR. The overall input third-order intercept point is found by adding half the input DR (refer back to the two-third rule, where half of two-third is one-third, so the upper level input is two-third of the intercept point).

Adding an AGC extends the DR of the receiver generally by extending the video and detection circuitry, as shown in Table 3.5. To determine the required AGC, two main factors need to be considered: the DR of the ADC and the input variations of the received signals.

Table 3.4 Receiver dynamic range calculations

Receiver	Gain/Losses	Sat. (dBm)	Noise (dBm)	IDR (dB)	DR (dB)
RF BW (kHz) (kTB)	100.00		−124		
RF components	2.00	50.00	−124	174.00	174.00
LNA noise figure (int.)	3.00		−121.00		
LNA (out)	30	7	−91.00	98.00	98.00
Filter (out)	−2	50	−93.00	143.00	143.00
Mixer (out)	−10	0	−103.00	103.00	103.00
IF Amp (out)	50	7	−53.00	60.00	60.00
IF bandwidth (kHz)	10		−63.00		
IF filter loss	−2	50	−65.00	115.00	115.00
Baseband mixer (out)	−10	0	−75.00	75.00	75.00
A/D converter (bits)	8	−27	−75.00	48.00	48.00

Table 3.5 Receiver dynamic range calculations using feedback AGC

Receiver	**Gain/Losses**	**Sat. (dBm)**	**Noise (dBm)**	**IDR (dB)**	**DR (dB)**
RF components	2.00	50.00	−174	224.00	224.00
RF BW (kHz)(kTB)	100.00		−124		
LNA noise figure (int.)	3.00		−121.00		
LNA (out)	30	7	−91.00	98.00	98.00
Filter (out)	−2	50	−93.00	143.00	143.00
Mixer (out)	−10	0	−103.00	103.00	103.00
AGC	**30**				
IF Amp (out)	50	7	−53.00	**60.00**	**90.00**
IF bandwidth (kHz)	10		−63.00		
IF filter loss	−2	50	−65.00	**115.00**	**145.00**
Baseband mixer (out)	−10	0	−75.00	**75.00**	**105.00**
A/D converter (bits)	8	−27	−75.00	**48.00**	**78.00**

Improves overall dynamic range DR. Does not improve instantaneous dynamic range IDR.

The DR of the ADC is dependent on the number of bits in the ADC. For an 8-bit ADC, the usable DR with a SNR of 15 dB is approximately 30 dB (theoretical is approximately 6 dB/bit = 48 dB). This establishes the range of signal levels that can be processed effectively. If the input variations of the received signals are greater than this range, then an AGC is required. If the IDR needs to be greater than the ADC range, then a feed-forward AGC needs to be implemented to increase IDR. A feed-forward AGC uses a part of the input signal to adjust the gain before the signal is received at the AGC amplifier, so there is no time associated with the response of the gain. The amount of feed-forward AGC required is calculated by subtracting the ADC range from the input signal range.

The feedback AGC extends the DR but does nothing to the IDR, as shown in Table 3.5. The IDR tests the range in which the receiver can handle an instantaneous change (much quicker than the AGC response time) in amplitude on the input without saturation or loss of detection.

- Does not improve IDR
- Improves overall DR

The IDR can be extended by using a feed-forward AGC or a log amplifier. The problem with a log amplifier is that the IDR is extended as far as amplitude, but since it is nonlinear the frequency IDR may not improve and might even worsen. The feed-forward AGC is faster than the signal traveling through the receiver. Therefore, the IDR is improved for both amplitude and frequency and can be done linearly.

DR is critical in analyzing receivers of all types. Depending on the particular design criteria, both amplitude and frequency DRs are important in the analysis of a receiver. Methods for determining DR need to be incorporated in the design and analysis of all types of receivers. This is essential in optimizing the design and calculating the minimizing factor or device that is limiting the overall receiver DR.

Graphical methods for calculating DR provide the engineer with a useful tool for analyzing receiver design. Both the DR and the IDR need to be considered for optimal performance. AGC can improve the receiver's DR tremendously but generally does not affect IDR. Feed-forward AGC and a log amplifier improve the IDR. However, the log amplifier is nonlinear and only improves the amplitude IDR and actually degrades the frequency IDR.

3.23 Sensitivity

Minimum Discernable Signal MDS is used to determine what the smallest signal level is needed for detection by the receiver. This is shown on an oscilloscope display in Figure 3.19. MDS is hard to measure exactly and is often times disputed however, it is a good method to use for approximation. Another method that is used for pulse measurements is known as TSS (Figure 3.20). The signal power is

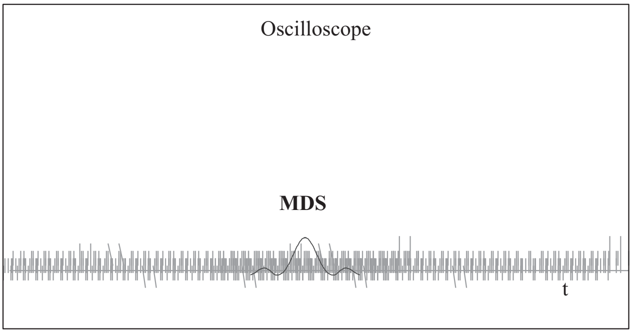


Figure 3.19 Minimum discernable signal MDS

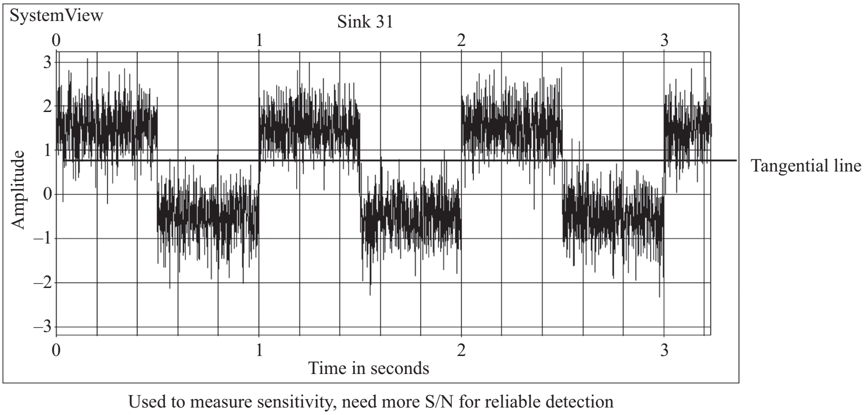


Figure 3.20 Tangential sensitivity (TSS) for a pulse system

increased until the bottom of the pulse noise is equal to the top of the noise level in the absence of a pulse. A tangential line is drawn to show that the pulse noise is tangential to the rest of the noise. With this type of measurement, a comparison between receivers can be made. The TSS measurement is good for approximately ± 1 dB.

SNR is usually specified to help define the MDS. For pulse systems, the TSS is utilized to improve the measurement of MDS and to compare receivers.

3.24 Digital signal processor

Once the analog signal is converted by the ADC to a digital signal, the digital signal processor finishes the reception by interpreting the data. Many receivers use DSP technology to accomplish most of the detection function. The RF/analog portion of the receiver is used only to downconvert the signal to a lower IF or baseband, spectrally shape the signal, and then convert it to the digital domain, where the remaining processing is digital. This enables the receiver to be configured for many receive applications and configurations for different waveforms by simply changing the software that drives the digital receiver.

DSP can do everything that analog processing can do, with the only limitation being the processor throughput required to sample fast enough for the frequency content of the signal. In conclusion, the sooner the signal can be processed digitally the better.

3.25 Summary

The receiver is an important element in the transceiver design. The receiver accepts the signal in space from the transmitter and amplifies the signal to a level necessary for detection. The LNA is the main contributor to the NF of the system. The superheterodyne is the most used receiver type and provides the most versatility by being able to apply a common IF. Saturation, compression, sensitivity, DR, reduction in unwanted spurious signals, and maximization of the SNR are the main concerns in designing the receiver. Mixers perform the downconversion process using spur analysis and selecting the correct mixer for the application. Two types of DR include amplitude, which is the most common way to express DR, and frequency DR, related to the two-tone third-order intercept point. Group delay plays an important role in digital communications, and careful consideration in the design will help reduce dispersion and ISI. Digital receivers perform most of the detection in data links today, and the ADC is used to translate an analog signal into the digital domain. The sooner the signal is in the digital domain, the better the receiver can optimize the detection process. Phase noise characteristics and definitions, aliasing and sampling theorem, image frequency and the importance of filtering are also included in this chapter.

3.26 Problems

1. What is a superheterodyne receiver?
2. What is placed in the transceiver that prevents the transmitted signal from entering the receiver on transmit?
3. What is the gain required by the receiver for a system that has an 8-bit ADC, maximum of 1 V, NF of 3 dB, bandwidth of 10 MHz, and source noise of 4 dB?
4. What is the third-order SFDR of the system in problem 2 with a +20 dBm *IP*3?
5. What is the usable DR of the system in problems 2 and 3 with a required SNR of 10 dB?
6. What is the NF without using an LNA with an IF NF of 3 dB and the total loss before the IF of 10 dB?
7. Calculate all intermodulation products up to third order for 10 and 12 MHz signals.
8. What is the shape factor for a filter with 0 dBm at 90 MHz, −3 dBm at 100 MHz, and −60 dBm at 120 MHz?
9. Find the required sample rate to satisfy the Nyquist criteria if the highest frequency component is 1 MHz?
10. What would be the maximum amplitude and phase error on a linear ADC using 2 bits of resolution?
11. What are the advantages and disadvantages for oversampling a received signal?
12. Why is constant group delay important in digital communications? What is the result if the group delay is not constant?
13. What does LNA stand for? Why is it important?
14. What happens if a system samples the input signal at less than the Nyquist rate for the highest frequency content?
15. Why is each bit of an ADC equal to 6 dB?

Further reading

Tsui, J. B. *Microwave Receivers with Electronic Warfare Applications*. Edison, NJ: SciTech Publishing, 2005.

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Chapter 4

AGC design and PLL comparison

Automatic gain control (AGC) is used in a receiver to vary the gain to increase the dynamic range (DR) of the system. AGC also helps deliver a constant amplitude signal to the detectors with different radio frequency (RF) signal amplitude inputs to the receiver. AGC can be implemented in the RF section, the intermediate frequency (IF) section, in both the RF and IF portions of the receiver, or in the digital signal processing (DSP) circuits. Digital AGCs can be used in conjunction with RF and IF AGCs. Most often, the gain control device is placed in the IF section of the receiver, but placement depends on the portion of the receiver that limits the DR. The detection of the signal level is usually carried out in the IF section before the analog-to-digital converter or analog detection circuits. Often the detection occurs in the DSP circuitry and is fed back to the analog gain control block. The phase-locked loop (PLL) is analyzed and compared with the AGC analysis, since both processes incorporate feedback techniques that can be evaluated using control system theory. The similarities and differences are discussed in the analysis. The PLL analysis is used only for tracking conditions and not for capturing the frequency or when the PLL is unlocked.

4.1 AGC design

An AGC adjusts the gain of the receiver automatically. It is used to increase the DR of the receiver by adjusting the gain depending on the signal level; large signals need less gain, small signals need more gain.

There are different types of AGC designs, the most common of which is a feedback AGC, where the signal level is detected and processed, and a signal is sent back to the AGC amplifier or attenuator to control the gain. It uses the feedback of the received signal to adjust the gain. Another type of AGC is known as a feed-forward AGC. This type of AGC detects the signal level and adjusts the gain farther down the receiver chain. This type of AGC uses feed-forward, not signal feedback and is used for high-speed control. Another type of control is dependent on time. This type of control is called a sensitivity time constant and is used in radar systems. For example, the change of gain is proportional to $1/t^2$ or other criteria that are dependent on the desired output and response. These types of gain control methods are shown in Figure 4.1.

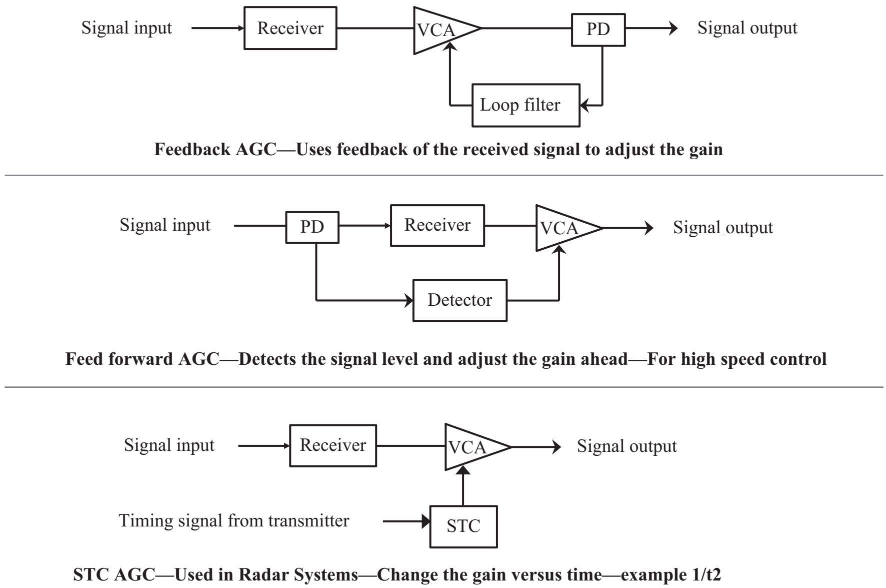


Figure 4.1 Different types of AGCs used in receivers

For the feedback AGC, the RF signal is downconverted and amplified, and the output is split and detected for use in the AGC to adjust the gain in the IF amplifiers. A voltage-controlled attenuator or variable gain amplifier, plus a linearizer, may be used for the actual control of the signal amplitude. The AGC voltage can be used to display the received power with an accuracy of about 0.5 dB by translating AGC volts to a received power equivalent which can be shown on a display or monitored by a personal computer.

The requirements for AGC in a system are established by using several parameters to determine the amount of expected power fluctuations that might occur during normal operation. Some of the parameters that are generally considered include the variation in distance of operation, propagation loss variations such as weather and fading, multipath fluctuations, variations in the antenna pattern producing a variation in the gain/loss of the antenna in comparison to a true omnidirectional antenna, and expected power fluctuations because of hardware variations in both the transmitter and the receiver.

A typical AGC used in receivers consists of an AGC amplifier and a feedback voltage to control the gain. A voltage-controlled attenuator can be used in place of the AGC amplifier; however, the noise figure of the receiver can be changed if the attenuation is large. This feedback system can be designed using basic control system theory.

The basic model is shown in Figure 4.2(a). This model is set up with the maximum gain from the AGC amplifier taken out of the loop as a constant, and the feedback loop subtracts gain from this constant depending on the detected signal level.

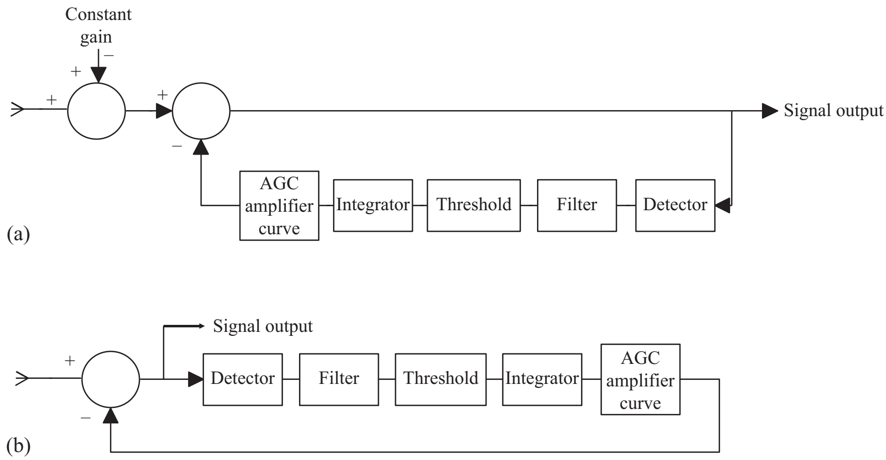


Figure 4.2 AGC block diagrams used in the control-system analysis. (a) Control system for the AGC and (b) control-system analysis for the second order AGC closed loop response

The feedback system is redrawn to show the point at which the stability analysis is done (Figure 4.2(b)). This determines the stability of the AGC loop, and the feedback gain is then equal to unity, which makes it more convenient to perform the analysis using feedback control theory.

4.2 AGC amplifier curve

The first step in designing an AGC is to determine the amount of gain control needed. This may be given directly, or it can be calculated given the receiver's range of operation. Sometimes, a compression ratio (range of the signal in versus range of the signal out) is provided. This will determine which AGC amplifier is needed.

Once the AGC amplifier is selected, the gain (dB) versus the control voltage (V) slope of the AGC amplifier is determined. The best way to do this is actual measurements in the lab, although some companies will furnish this curve in their data sheets. This slope is generally nonlinear. The AGC amplifier should be chosen so that the gain versus control voltage is as linear as possible, allowing the loop gain and noise bandwidth to remain stable as the control voltage is changed. An estimated linear slope can be applied depending on the design criteria. An example of an amplifier curve and an estimated linear slope is shown in Figure 4.3.

The slope is in dB/V and is used in the loop gain analysis. The estimated linear slope should be chosen where the AGC is operating most of the time. However, the AGC system should be designed so that it does not become unstable across the

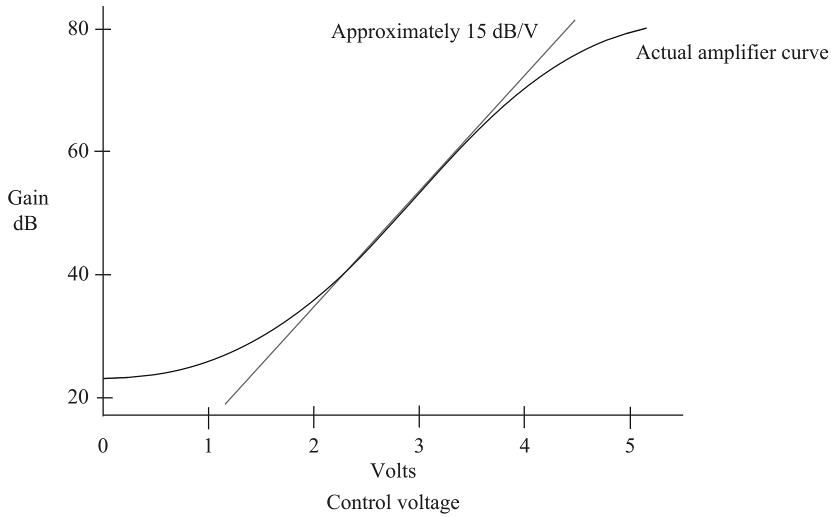


Figure 4.3 Typical gain versus control voltage for an AGC amplifier

entire range of operation. The gain constant should be chosen at the steepest portion of the slope so that the loop gain will always be less than this gain value. This will cause the response time outside this approximation slope of the AGC to be longer (smaller bandwidth), which means the noise in the loop will always be less. Therefore, since the bandwidth can only get smaller, the reduction of desired modulation will always be smaller. This prevents the loop from becoming unstable. Unless a slower response time presents a problem (as in the case of fading when a very slow AGC is used), these design criteria can be used in most situations.

4.3 Linearizers

Linearizers can be used to compensate for the amplifier's nonlinearity and create a more constant loop gain response over the range of the amplifier. A linearizer is a circuit that produces a curve response that compensates the nonlinear curve such that the sum of the two curves results in a linear curve (Figure 4.4). Note that the linear response is on a linear (V) versus log (P) scale.

A linearizer can be designed by switching in diodes in a piecewise fashion, which alters the gain slope to compensate for the AGC amplifier gain curve. As the diodes are switched on, they provide a natural smoothing of the piecewise slopes due to the slope of the diode curve itself. This allows for fairly accurate curve estimations. Note that in Figure 4.4, the signs of the slopes have been neglected for simplicity. Care needs to be taken to ensure the correct sense of the loop to provide negative feedback. Linearizers can also be created and used in the digital processor, which is more accurate and is more immune to temperature variations and changes to the slopes of the desired curves.

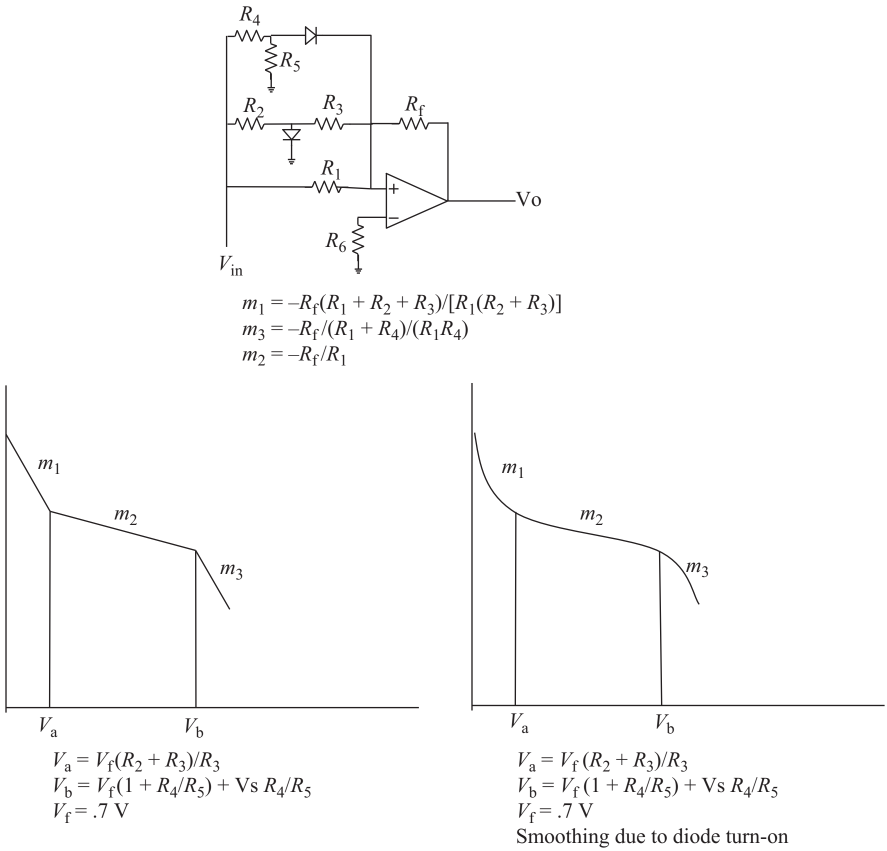


Figure 4.4 Linearizer circuit to compensate for a nonlinear response

4.4 Detector

The next step is the detector portion of the AGC. The sensitivity (gain) of the detector and the linearity of the curve (approaching a log of the magnitude-squared device) are important parameters in selecting the detector. The linear section of the detector can be chosen on the *square-law* portion of the curve (low-level signal) or the *linear* portion of the curve (high-level signal). The output power or voltage of the AGC is given or chosen for a particular receiver. The operating point can influence the choice of the detector used. The detector provides a voltage-out (V)/power-in (dBm) curve, or V/dB slope, which can be measured or given. The power-in is an actual power level and is given in dBm. The change of power-in levels is a gain or loss given in dB. Therefore, the slope is a change of power-in levels and is given as V/dB not V/dBm. The ideal detector needs to be linear with respect to:

$$V_o = \log\left[V_i^2\right] (\text{log square laws lope})$$

The slope can be calculated if the detector is ideal:

$$(V + dV) - (V - dV)/[2\log(V + dV)/V - 2\log(V - dV)/V]$$

where V is the operating voltage level of the detector and dV is the small variation of the operating voltage level.

Note that if the detector is linear with respect to:

$$V_o = \log[V_i](\log\ \text{slope})$$

The difference in the slope is a scale factor (2).

An amplifier stage with a gain of two following this detector would make it equal to the previous detector. A linearizer could be designed to compensate for either slope.

A linearizer design is shown in Figure 4.4. This method uses diode activation to set the break points of the curves, and the diode turn-on characteristics smooth the curves to a very close approximation of the desired curve.

The slope of a real detector is not linear. However, the slope is chosen so that it is linear in the region in which it is operating most of the time (Figure 4.5). This provides another part of the loop gain, given in V/dB.

The nonlinearity of the detector will change the instantaneous loop gain of the feedback system. Unless the system goes unstable, though, this is generally not a

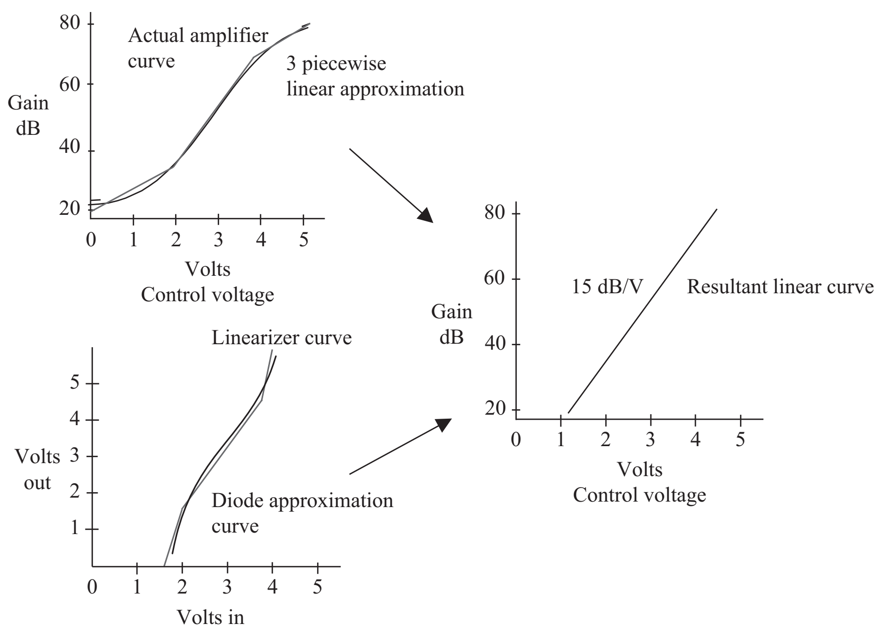


Figure 4.5 Using a linearizer curve to obtain a linear response for the amplifier control

problem because the operating level to the diode is held constant by the AGC. Since the diode will always be operating in steady state at one point on the curve, the importance of linearizer is not significant compared with the AGC amplifier linearizer.

A linearizer can be used to help create a more linear slope from the detector so that the loop gain is more constant. If the modulating signal is large compared with the nonlinearity of the diode curve, there could be some distortion of the modulating signal. Also, there might be some distortion when the signal is on the edge of the AGC range of operation.

The resistor–capacitor (RC) time constant following the diode should be much larger than the period of the carrier frequency and much smaller than the period of any desired modulating signal. If 1/RC of the diode detector approaches the loop time response, then it should be included in the loop analysis. If 1/RC is greater than an order of magnitude of the loop frequency response, then it can be ignored, which is generally the case.

Since an AGC is generally not attempting to recover a modulating signal, as in amplitude modulation (AM) detection, failure-to-follow distortion is not considered. However, if the rate of change in amplitude expected is known, then this can be used as the modulating signal in designing for the response time. If a modulating signal (like a conical scan system produces to achieve angle information) is present, the total AGC bandwidth should be at least 10 times smaller to prevent the AGC from substantially reducing the conical scan modulating frequency. Here again, this depends on the design constraints. The modulation is actually reduced by $1/(1 + \text{loop gain})$. This is the sensitivity of the loop to the frequency change of the input.

A standard diode can be used to detect the signal. This is a common silicon diode, and the response of the diode is shown in Figure 4.6. Notice that this diode

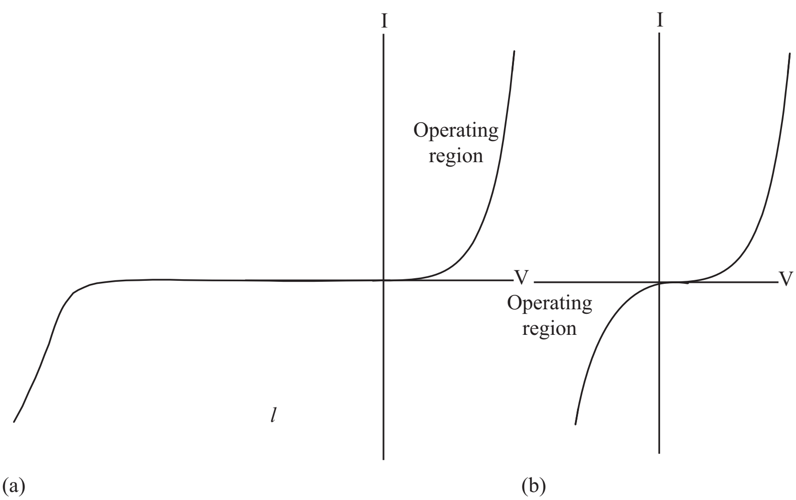


Figure 4.6 Slope of some typical detectors. (a) Standard diode and (b) back diode

needs to be biased on, since the turn-on voltage is approximately 0.7 V for a silicon diode. Another type of diode that can be used to detect the signal is called a back diode. It provides very good sensitivity as a detector and uses the negative (back-biased) portion of the curve for detection (Figure 4.6). This detector operates well in the presence of low-level signals and requires no bias voltage for the diode, since the operating region begins at zero.

The amplifier before the detector needs to be capable of driving the detector. A buffer amplifier can be used to supply the necessary current as well as to isolate the received signal from the loading effects of the detector. With a power divider and high-frequency detectors, this may not be a problem. The detector is matched to the system, and the AGC amplifier output can drive the detector directly through the power divider.

4.5 Loop filter

A loop filter needs to be designed to establish the frequency response of the loop and to stabilize the loop. The loop filter should follow the detector to prevent any high-frequency components from affecting the rest of the loop. A passive phase-lag network is used. The schematic consists of two resistors and a capacitor and is shown in Figure 4.7 along with the transfer function. This filter provides a pole and a zero to improve the stability of the loop. This filter is used to affect the attenuation and the roll-off point, since the phase shift for a phase lag filter generally has a destabilizing effect.

4.6 Threshold level

A threshold needs to be set to determine the power level output of the AGC amplifier. A voltage level threshold in the feedback loop is set by comparing a stable voltage level with the output of the loop filter. This sets the power level out of the AGC amplifier. A threshold level circuit consists of an operational amplifier with a direct current (DC) offset summed in (Figure 4.7). In this case, if the output of the loop filter is less than the offset, then the output of the threshold device will be negative. If the loop filter output is greater than the offset, the output will be positive. This determines which way the gain is adjusted. The offset is equal to $-(\mathrm{R}_5/\mathrm{R}_4)V_{\mathrm{DC}}$, as shown in Figure 4.7.

An additional gain is added to the loop and is selected for the particular AGC signal level and diode gain. This gain should be large enough to increase threshold sensitivity but small enough to prevent saturation of the loop amplifiers. The gain constant is defined as K_c and is equal to $-(\mathrm{R}_5/\mathrm{R}_6)V_1$, as shown in Figure 4.7.

4.7 Integrator

With just the phase-lag filter in the loop, the system is classified as a type-0 system, and there will be a steady-state error for a step response. If an integrator is included

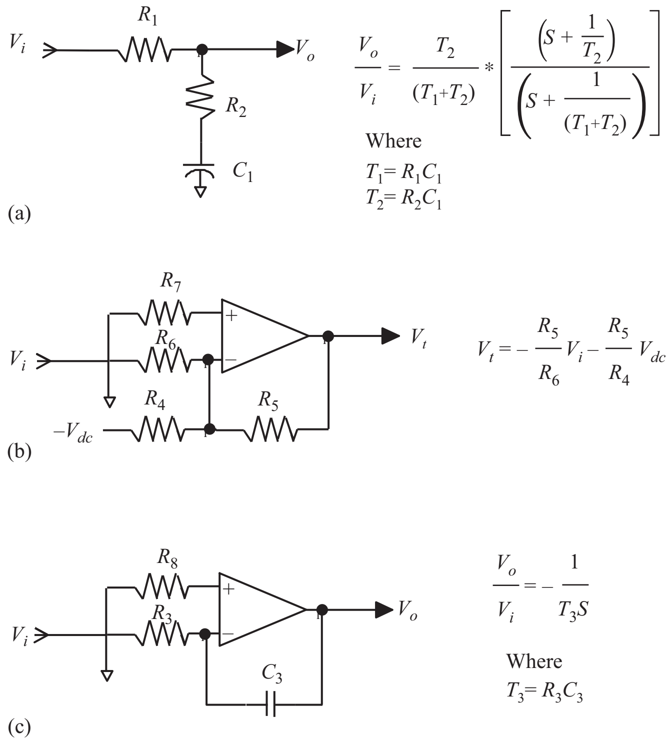


Figure 4.7 Loop filter components. (a) This is a lag filter providing a zero and a pole, (b) this is a threshold amplifier summing in the offset voltage which determines the AGC output level, and (c) this is an integrator to make the system a type 1 feedback system and ensures that the output of the AGC is exact, no steady-state error for a step response

in the feedback loop, then the system is a type-1 system and the steady-state error for a step response is zero. Therefore, an integrator is included in the loop. An integrator, using an operational amplifier, is shown in Figure 4.7. The output of the threshold device is integrated before controlling the gain of the AGC amplifier. With no signal present, the integrator will integrate to the rail of the operational amplifier (op amp) and stay there until a signal greater than the threshold is present. The AGC amplifier is held at maximum gain during this period of time, which is the desired setting for no signal or very small signals. The response time for the op amps to come out of saturation is negligible compared with the response time of the loop. If the voltage level out of the loop amplifiers is too large for the AGC amplifier, then a voltage divider or a voltage clamp should be used to protect

the AGC amplifier. The DC gain of the integrator should not be limited by putting a resistor in parallel with the feedback capacitor, or there will be a steady-state error. This makes the loop a second-order system and is easily characterized by using control system theory. An integrator should always be included in the AGC design in order to keep the signal level output of the AGC amplifier and into the detector at the exact desired level with no variations for a step change in received power.

Most systems today use DSP techniques to detect the signal level and then provide feedback to control the gain of the amplifier or adjust the attenuator value. This is a feedback loop with the potential to oscillate, so careful design using control theory analysis is needed. This analysis includes the time delays, frequency responses, and gains through the digital circuitry path.

4.8 Control theory analysis

The open-loop transfer function including the integrator is:

$$T_{\mathrm{sol}} = \frac{K_a K_d K_c T_2\left(S + \frac{1}{T_2}\right)}{T_3(T_1 + T_2)S\left(S + \frac{1}{T_1+T_2}\right)}$$

where:

$K_a =$ AGC amplifier loop gain
$K_d =$ detector loop gain
$K_c =$ gain constant
$T_1 = R_1 \times C_1$
$T_2 = R_2 \times C_1$
$T_3 = R_3 \times C_3$
$T_{\mathrm{sol}} =$ open-loop transfer function
$S = j\omega$ (assumes no loss).

The closed-loop response is

$$G(s)/(1 + G(s)) \text{ with } H(s) = 1$$

$$T_{\mathrm{scl}} = \frac{K\left(S + \frac{1}{T_2}\right)}{S^2 + \left(\frac{1}{T_1+T_2} + K\right)S + \frac{K}{T_2}}$$

where:

$K = K_a K_d K_c (1/T_3)(T_2/[T_1 + T_2])$
$T_{\mathrm{scl}} =$ closed-loop transfer function

From control system theory for the previous second-order system,

$$2\zeta\omega_n = K + 1/(T_1 + T_2)$$

and:

$$\omega_n^2 = K/T_2$$

where ω_n is the natural frequency of the system and ζ is the damping ratio.

The transfer functions and the block diagram of the analysis are shown in Figure 4.8.

The natural frequency (ω_n) should be set at least ten times smaller than any desired modulating frequency. The damping ratio is chosen so the system responds to a step function as fast as possible within a given percent of overshoot. With a 5% overshoot (PO), the minimum damping ratio is 0.707. This means the system is slightly underdamped. The damping ratio is chosen depending on the design criteria. If the overshoot is too high, the damping ratio should be changed. Once the damping ratio and the natural frequency are chosen for a system, the time constants can be solved mathematically.

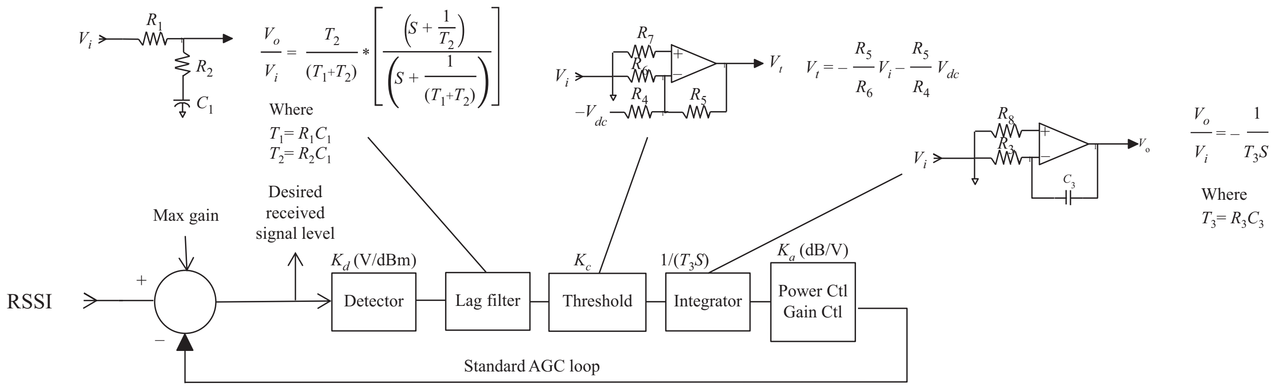
Suppose the conical scan information modulation frequency of a radar system is equal to 30 Hz. The loop frequency response needs to be an order of magnitude less so that it does not interfere with the positioning information. Therefore, the loop frequency response is less than 3 Hz so that the AGC does not track out the angle information. The following choices are made:

Choose $f_n = 1.6$ Hz $\omega_n = 10$ rad/s.

Choose $\zeta = 0.707$ PO $= 5\%$.

The poles and zeros for the root locus criteria are found by using $G(s)H(s)$. Note that $1/T_2$ determines the zero location for both the open-loop and closed-loop cases, since $H(s) = 1$. The zero is placed at $-2\zeta\omega_n$, and the root locus plot is shown in Figure 4.9. This moves the root locus, the migration of the poles, away from the $j\omega$ axis to prevent oscillatory responses with a change in gain. This also places the poles on the tangent of the root locus circle at the given damping ratio so that as the gain varies the damping ratio is always greater. One pole, due to the integrator, migrates from the $j\omega$ axis along the σ axis as the loop gain constant K increases. As K approaches zero, the pole due to the integrator lies on the $j\omega$ axis. This is something to keep in mind when designing with nonlinear devices. As the loop gain decreases, the integrator pole migrates closer to the $j\omega$ axis and becomes more oscillatory. However, a pole on the axis means that the system is overdamped $[\zeta > 1]$ and will not oscillate. The only way this system can become unstable is with stray poles close to the origin, which is unlikely. The loop gain K is chosen in this example, so the operating point lies on the root locus circle with the desired damping factor (Figure 4.9).

Examples of choosing different dampening factors are shown in Figure 4.10. The top graph shows an underdamped factor which contains oscillatory characteristics and eventually converges to the desired level. The middle graph shows a quick response time to the desired level with some overshoot. This is generally a good approach. The percentage is changed for the particular design. If the AGC control signal is too oscillatory, then the trade-off would be less oscillatory for



Basic second order AGC closed-loop response.

$$\text{Open-loop transfer function} = T_{\text{sol}} = G(s) = K_d * \frac{T_2}{(T_1+T_2)} \frac{\left(S+\frac{1}{T_2}\right)}{\left(S+\frac{1}{T_1+T_2}\right)} * K_c * \frac{1}{T_3S} * K_a = \frac{K_aK_dK_cT_2\left(S+\frac{1}{T_2}\right)}{T_3(T_1+T_2)S\left(S+\frac{1}{T_1+T_2}\right)}$$

$$= \frac{K\left(S+\frac{1}{T_2}\right)}{S\left(S+\frac{1}{T_1+T_2}\right)} \qquad K = \frac{K_aK_dK_cT_2}{T_3(T_1+T_2)}$$

$$\text{Closed-loop transfer function} = T_{\text{scl}} = G(s) / [1 + G(s)H(s)] = \frac{K\left(S+\frac{1}{T_2}\right)}{S^2 + \left(\frac{1}{T_1+T_2} + K\right)S + \frac{K}{T_2}}$$

Figure 4.8 Transfer functions and block diagram used in the analysis of the AGC

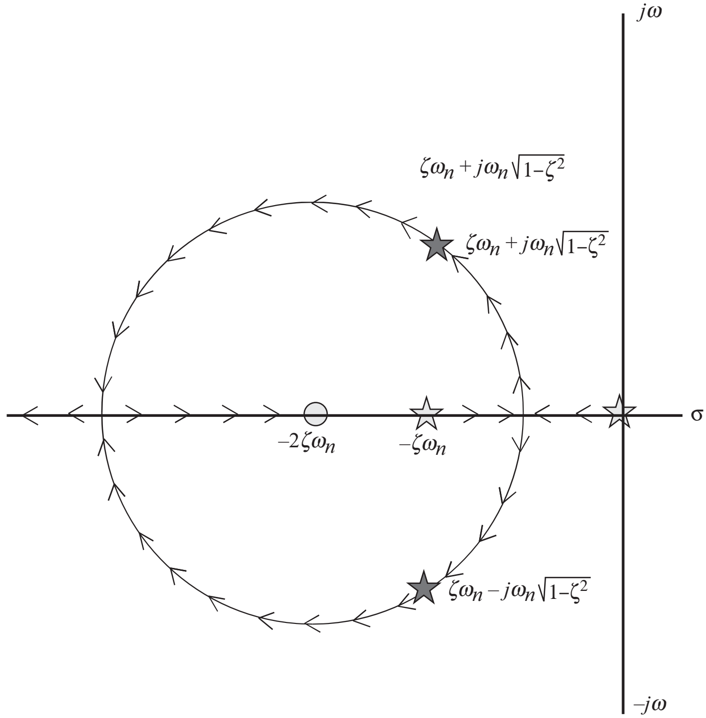


Figure 4.9 Root locus plot for AGC analysis

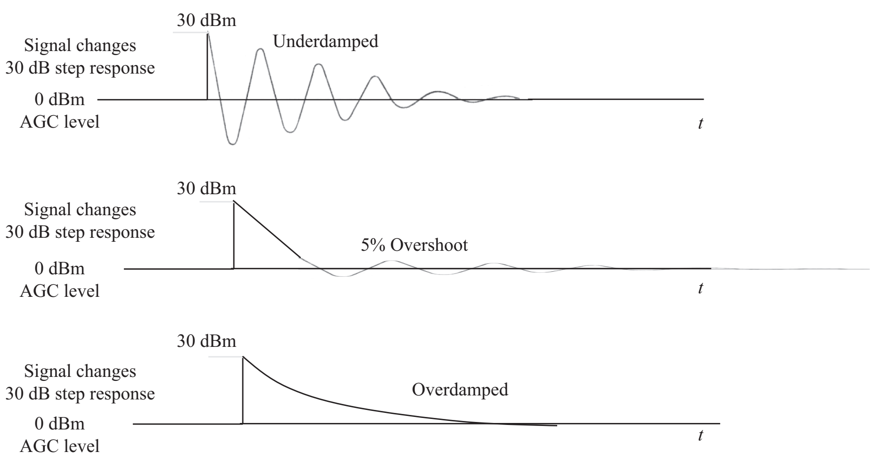


Figure 4.10 AGC responses for different dampening factors

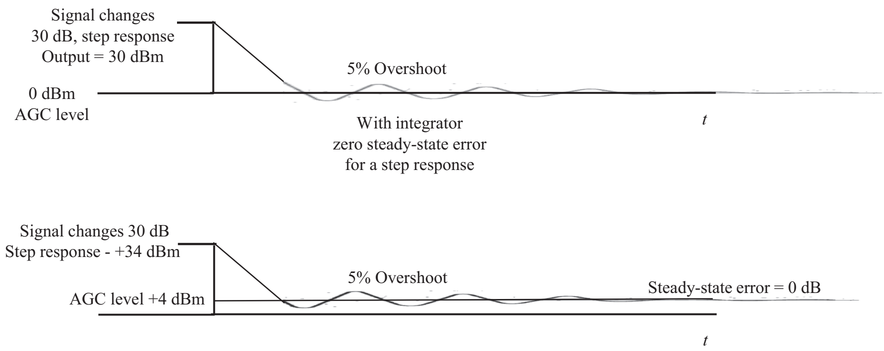


Figure 4.11 Steady-state error analysis

slower convergent to the desired level signal. If time is critical for convergence, then the damping factor would be chosen for more oscillatory behavior. The bottom graph is overdamped with no oscillatory behavior but a long convergent time (Figure 4.10).

The steady-state error can be chosen for the level desired. The threshold is changed for the desired output and the same convergent curves apply (Figure 4.11). The top graph shows a convergent to 0 dBm, and the bottom shows the convergent to +4 dBm, with both of the curves having the same overshoot and both converging to 0 dB steady-state error.

4.8.1 AGC design example

An example for an AGC design is shown in Table 4.1. The parameters for this example are:

$\omega_n = 10$
$\zeta = 0.707$
$K_a = 15$
$K_d = 0.15$
$K_c = 6.27$

Choose the zero location:

$$T_2 = 1/(2\zeta\omega_n) = 1/(14.14)) = 0.0707$$

Using the previous root locus equations:

$$K = T_2\omega_n^2 = 0.0707(10)^2 = 7.07$$
$$2\zeta\omega_n = K + 1/(T_1 + T_2)$$

Solving for T_1:

$$T_1 = 1/(2\zeta\omega_n - K) - T_2 = 1/(14.14 - 7.07) - 0.0707 = 0.0707$$

Table 4.1 An example of an AGC design

2ND ORDER AGC DESIGN:											
Enter Constants:						Tclosed loop = [K(S + 1/T2)]/[S*S + (1/(T1 + T2) + K)S + K/T2]					
Choose damping ratio z =			0.707								
Choose natural frequency fn =			10	Hz							
AGC amplifier loop gain = Ka =			15	dB/Volt							
Detector Loop Gain = Kd =			0.15	Volt/dB		Percent Overshoot =		4.325493	%		
Other component Gain = Kc =			10	Volt/Volt		wn =	62.83185	rad/sec			
Calculations:						**Detector Design:**					
The zero is set at 2zwn			T2 =	0.011256	sec	Enter RF source impedance =			5000	ohms	
K = T2*wn*wn =		44.43554				Detector gain =		0.15	Volts/dB		
2zwn = K+1/(T1+T2)											
Solve for T1 = 1/(2zwn-K)-T2			T1 =	0.011262	sec	Detector time constant should be much faster than the AGC					
K = KaKdKc(1/T3)(T2/[T1+T2])						response time of the feedback loop or it needs to be included					
T3 = KaKdKc(T2/[T1 + T2])/K			T3 =	0.253099	sec	and much slower than the RF frequency.					
Enter Values:						RF frequency =		0.05	MHz		
Choose C1 =		1	uf			Time constant =		200	us		
Choose C2 =		1	uf			Choose C2 =		0.1	uf		
Choose C3 =		1	uf			R9 =		5000	ohms		
						R10 =		2	kohms		
Calculations:											
R1 =	11.26246	kohms				**Offset Design:**					
R2 =	11.25565	kohms				Choose dBm out =		5	dBm		
R3 =	253.0992	kohms				Enter coupler loss =		3	dB		
						Detector input =		2	dBm		
Etc.						Detector input =		1.407521	Vpeak		
R7 =	3.005689	kohms				Detector output =		0.995268	Vrms		
R8 =	126.5496	kohms									
						Amplifier output =		9.952679			
Distortion on Signal:						Choose R5 =		5	kohms		
Choose freq. of operation =			25	kHz		Choose Vdc =		15	volts		
Frequency in radians =			157079.6			R6 =		0.5	kohms		
Tclosed loop = [K(S + 1/T2)]/[S*S + (1/(T1 + T2) + K)S + K/T2]						R4 =		7.535659	kohms		
Numerator Real =		3947.842									
Numerator Imag =		6979918									
Denominator Real =		-2.5E+10									
Denominator Imag =		20928878									
Closed Loop Mag =		0.000283									
% affecting signal =		0.028289	%								

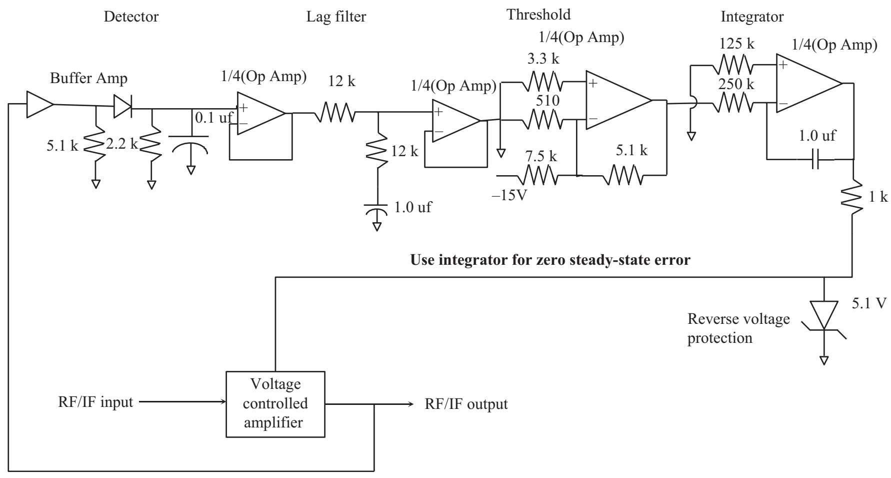


Figure 4.12 A block diagram of the AGC design example

Using the total loop gain expression to solve for T_3:

$$K = K_a K_d K_c(1/T_3)(T_2/[T_1 + T_2])$$
$$T_3 = K_a K_d K_c(T_2/[T_1 + T_2])/K$$
$$= 15(0.15)(6.27)[(0.0707)/(0.0707 + 0.0707)]/7.07 = 1.0$$

Once the time constants are solved, the selection of the resistors and capacitors is the next step. Low-leakage capacitors are preferred, and the resistors should be kept small. This is to prevent current limiting of the source by keeping the current large compared with the current leakage of the capacitor. There are trade-offs here, depending on parts available and values needed.

Choose $C_1 = C_3 = 1\ \mu F$
$T_1 = R_1C_1$, therefore $R_1 = T_1/C_1 = 0.0707/1\ \mu F = 70.7\ k\Omega$
$T_2 = R_2C_1$, therefore $R_2 = T_2/C_1 = 0.0707/1\ \mu F = 70.7\ k\Omega$
$T_3 = R_3C_3$, therefore $R_3 = T_3/C_3 = 1.0/1\ \mu F = 1\ M\Omega$

A block diagram showing the design of the AGC is shown in Figure 4.12.

4.9 Modulation frequency distortion

The reduction of the modulating frequency (e.g., the Conscan frequency information in the preceding example) is calculated by solving the closed-loop transfer function for the modulating frequency ω_m.

$$\omega_m = 188.5\ \text{rad/s}$$
$$T_{\text{closed loop}} = 7.05(j188.5 + 14.14)/[(j188.5)^2 + 14.14(j188.5) + 100]$$
$$= 100 + j1329/(-35,432 + j2665) = 0.04/\!-90^\circ.$$

The 0.04 number means that for an input gain change of a signal level at the modulating frequency ω_m, the AGC amplifier will change its gain by 0.04 of the input gain change. The modulating signal is reduced by 0.04 times the input gain change in dB.

For example, if the input signal level changes from 0 to 10 dBm at the modulating frequency, the input gain change is 10 dB. The feedback loop response reduces the 10 dB change by a magnitude of, say, 0.04 or 0.4 dB. This means the AGC amplifier changes gain by 0.4 dB for an input change of 10 dB. However, since there is a phase shift of the feedback signal, the amplitude reduction will be less, since it is associated with the cosine of the angle. In equation form:

$$\text{Signal input} = P_{\text{in dBm}} + A\cos(\omega_m t)_{\text{dB}}$$
$$\text{Signal output} = P_{\text{in dBm}} + A\cos(\omega_m t)_{\text{dB}} - HA\cos(\omega_m t + C)_{\text{dB}}$$

where A is the amplitude of the modulating signal (conscan), H is the amplitude through the feedback loop, C is the phase shift through the loop, and t is the time.

The superposition of the two cosine waves in the output equation results in the final gain (in dB) that is added to the power. We can also take the change in dB caused by the input modulation amplitude and phase, subtract the loop input and phase by converting to rectangular coordinates, and then convert back to polar coordinates. For example,

Input = 10 dB angle = 0°
Loop response = mag(0.04 × 10) = 0.4 dB angle (−90° + 0°) = −90
Total response = $(10 + j0) - (0 - j0.4) = 10 + j0.4 = 10.008$ dB at an angle of 2.3°

The magnitude changed by 0.008 dB and the phase of the modulating signal shifted by 2.3°.

Note that this assumes that the change in gain is within the DR of the AGC. If the gain change is outside of the AGC DR, then the magnitude of the loop response is 0.04 times only the portion of the AGC range that is being used.

To determine the resultant amplitude and phase for the general case:

$$H(\omega) = [H]e^{jC} = [H]\cos C + j[H]\sin C$$

Thus,

$$\{1 - H(\omega)\} = 1 - \{[H]\cos C - j[H]\sin C\}$$

The magnitude equals

$$\text{mag}[1 - H(\omega)] = \sqrt{\left\{1 + [H]^2 - 2[H]\cos C\right\}}$$

The angle equals

$$\tan^{-t}\{-[H]\sin C/(1 - [H]\cos C)\}$$

4.10 Comparison of the PLL and AGC using feedback analysis techniques

PLLs are important elements in the design of synthesizers and carrier recovery circuits. They are used in almost all receiver designs and have become a basic building block design tool for the engineer. This section simplifies the analysis of PLL stability and makes a comparison between AGC and the PLL.

4.11 Basic PLL

The basic PLL standard block diagram is shown in Figure 4.13. The operation of the PLL takes the input frequency and multiplies this frequency by the voltage-controlled oscillator (VCO) output frequency. If the frequencies are the same, then a phase error is produced. The nature of the PLL forces the phase of the VCO to be in quadrature with the phase of the incoming signal. This produces the sine of the phase error:

$$\cos(\omega t + a)\sin(\omega t + b) = 1/2[\sin(a - b) + \sin(2\omega t + a + b)]$$

Using a low-pass filter to eliminate the sum term produces

$$1/2[\sin(a - b)]$$

if $a = b$, then

$$1/2[\sin(0)] = 0$$

The zero-voltage output sets the VCO to the correct steady-state value. If the voltage changes due to a change in input phase, the VCO changes until the phase

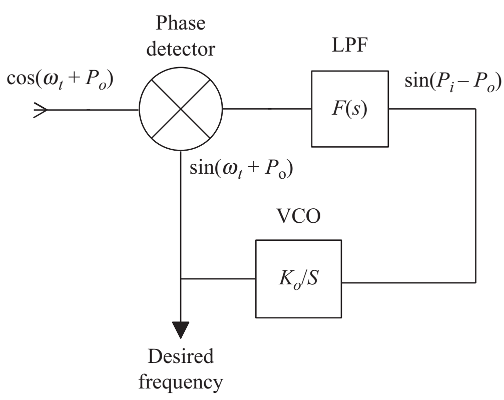


Figure 4.13 Basic diagram of a phase lock loop

error is approximately equal to zero. The low-pass filter removes higher frequency components and also helps to establish the loop gain. The PLL tracks a phase change over time, and since the change of phase per unit time is frequency, the PLL tracks frequency. Since it performs this task by converting the changing phase into a changing voltage that controls the VCO, the analysis is performed using phase as the parameter.

4.12 Control-system analysis

PLLs can be designed using basic control system theory, since it is a feedback system. The analysis is very closely related to the AGC study. To analyze the stability of the PLL, the basic diagram is redrawn (Figure 4.14).

The block diagram for the PLL is almost identical to the AGC block diagram. The VCO contains an integrator and a gain constant, whereas the AGC amplifier only has a gain constant and the integrator is added as another block. The integrator included in the VCO produces a type-1 system, so for a step change in phase, the PLL will have a steady-state error of zero. For the AGC, the added integrator provides a steady-state error of zero for a step change (in dB) of input power. If a zero steady-state error for a step in frequency is desired, another integrator needs to be included in the PLL. However, this chapter will limit the analysis to a second-order type-1 system.

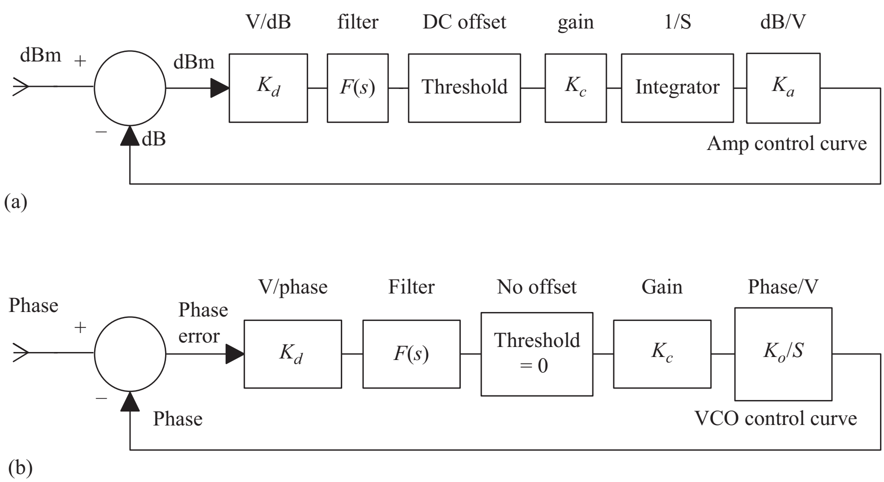


Figure 4.14 Comparison between AGC and PLL feedback systems. (a) AGC control-system analysis block diagram and (b) PLL control-system analysis block diagram

4.13 Detector

The phase detector or mixer in the PLL performs two operations (Figure 4.14). First, the detector takes the input phase and subtracts the feedback phase from the VCO to produce a phase error. Second, the detector converts the phase into voltage with a particular slope depending on the phase difference and the frequency of operation. This process for the AGC uses power (in dBm) for the input and subtracts the gain (in dB) to produce a power level error that is converted to voltage. The AGC detector's operation point is set at the desired dBm output, so the detector gain is the same around the operation point regardless of the input operating power. If the desired frequency is set for the PLL, then the same criteria apply. However, in many cases, when a PLL is used, the desired operating frequency changes (as is the case for synthesizer design). Thus, the detector gain becomes a function of the input frequency or the input operating point. Care must be taken when selecting the gain constant for the VCO detector because it depends on the input frequency. There are approaches for linearizing the detector, but for this analysis, the slope is chosen as a constant. If the frequency is constant, then the slope is easily chosen, and the linearity is generally not a problem.

4.14 Loop filter

The passive phase-lag network used for the loop stabilizing filter is identical for both the AGC and the PLL. The actual numerical values will be different depending on the loop requirements. The loop filter should follow the detector to prevent any high-frequency components from affecting the rest of the loop. The schematic consists of two resistors and a capacitor and is shown in Figure 4.7 along with the transfer function. This circuit provides a pole and a zero to improve the stability of the loop. The filter is used to affect the attenuation and the roll-off point, since the phase shift for a phase-lag network generally has a destabilizing effect.

4.15 Loop gain constant

Additional gain is added to the loop and is selected for the particular operating levels of the PLL. The gain constant is defined as K_c and is equal to $-R_5/R_6$ (Figure 4.7). The gain constant is used in both the PLL and the AGC cases and are different depending on each of their proper operating points. An added threshold is not required in the PLL, since the desired phase difference is always zero, regardless of the desired frequency of operation.

4.16 Integrator

The integrator for the PLL is inherent in the loop, whereas the AGC integrator is added to the loop. The reason for this is because the PLL operates using phase as

the parameter and the VCO actually delivers frequency. Therefore, the PLL forces the VCO to deliver phase that inherently adds an integrator because the integral of frequency is phase.

4.17 Conversion gain constant

The conversion in the AGC circuitry from volts to dBm produces a slope gain constant (dB/V) and is chosen considering the trade-off between stability and response time. The conversion in the PLL is a two-step process. First, the slope conversion constant converts from voltage to frequency, which is the constant for the VCO labeled K_o. Second, the inherent integration converts frequency into phase. The constant, K_o, for the PLL is not linear. This is also true for K_a in the AGC analysis because a linear approximation is used. A linearizer can be implemented for either the PLL or the AGC circuits.

4.18 Control theory analysis

The open-loop transfer function for the PLL is identical to the open-loop transfer function for the AGC except that the integrator is considered $1/S$ instead of $1/(T_3S)$ as follows:

Open-loop transfer function for AGC:

$$T_{\text{sol}} = \frac{K_a K_d K_c T_2 \left(S + \frac{1}{T_2}\right)}{T_3(T_1 + T_2)S\left(S + \frac{1}{T_1+T_2}\right)}$$

Open-loop transfer function for PLL:

$$T_{\text{sol}} = \frac{K_o K_d K_c T_2 \left(S + \frac{1}{T_2}\right)}{(T_1 + T_2)S\left(S + \frac{1}{T_1+T_2}\right)}$$

where:

$K_a =$ AGC amplifier loop gain
$K_o =$ VCO gain
$K_d =$ detector loop gain
$K_c =$ gain constant
$T_1 = R_1 \times C_1$
$T_2 = R_2 \times C_1$
$T_3 = R_3 \times C_3$
$T_{\text{sol}} =$ open-loop transfer function
$S = j\omega$ (for a lossless system)

The closed-loop response is:

$$T_{\mathrm{scl}} = \frac{K\left(S + \frac{1}{T_2}\right)}{S^2 + \left(\frac{1}{T_1 + T_2} + K\right)S + \frac{K}{T_2}}$$

where:

$K = K_a K_d K_c (1/T_3)(T_2/[T_1 + T_2])$ for the AGC
$K = K_o K_d K_c (T_2/[T_1 + T_2])$ for the PLL
T_{scl} = closed-loop transfer function for both the AGC and PLL

From control system theory for the previously given second-order system,

$$2\zeta\omega_n = K + 1/(T_1 + T_2)$$

and

$$\omega_n^2 = K/T_2$$

where ω_n is the natural frequency of the system and ζ is the damping ratio.

This analysis applies to the condition of the loop being already phase locked and not capturing the signal.

The loop gain for the PLL is selected considering several factors:

- Stability of the loop
- Lock range and capture range
- Noise in the loop
- Tracking error

The wider the bandwidth, the larger the lock and capture ranges and also the more noise in the loop. The lock range is equal to the DC loop gain. The capture range increases with loop gain and is always less than the lock range. If the maximum frequency deviation or the maximum phase error desired is given, then the open-loop gain can be established. When designing a synthesizer, the crossover between the crystal reference and the VCO with respect to phase noise is chosen for best noise performance for the loop bandwidth. Once the frequency is chosen, the same stability analysis for the PLL can be performed as was done for the AGC circuit analysis.

The damping ratio is chosen so the system responds to a step function as fast as possible within a given percentage of overshoot. With a 5% overshoot, the minimum damping ratio is 0.707. This means the system is slightly underdamped. The damping ratio is chosen depending on the design criteria. If the overshoot is too high, the damping ratio should be changed. Once the damping ratio and the natural frequency are chosen for a system, the time constants can be determined mathematically. An example is shown as follows:

Determine loop bandwidth = 160 Hz
Choose $\omega_n = 1.0$ krad/s $f_n = 160$ Hz
Choose $\zeta = 0.707$ PO = 5%

The poles and zeros for the root locus criteria are found by using $G(s)H(s)$. Note that $1/T_2$ determines the zero location for both the open-loop and closed-loop cases, since $H(s) = 1$. The zero is placed at $-2\zeta\omega_n$, and the root locus plot is shown in Figure 4.9. This moves the root locus, the migration of the poles, away from the $j\omega$ axis to prevent oscillatory responses with a change in gain. It also places the poles on the tangent of the root locus circle at the given damping ratio so that the damping ratio is always greater as the gain varies. Due to the integrator, one pole migrates from the $j\omega$ axis along the σ axis as the loop gain constant K increases. As K approaches zero, the pole due to the integrator lies on the $j\omega$ axis. As the loop gain decreases, the integrator pole migrates closer to the $j\omega$ axis and becomes more oscillatory. However, a pole on the axis means that the system is overdamped ($\zeta > 1$) and will not oscillate. The only way this system can become unstable is with stray poles close to the origin. The loop gain (K) is chosen in this example so that the operating point lies on the root locus circle with the desired damping factor (Figure 4.9).

For the example, choose T_2:

$$T_2 = 1/(2\zeta\omega_n) = 1/1.414k = 707.2 \times 10^{-6}$$

Using the previous root locus equations:

$$\begin{aligned} K &= T_2\omega_n^2 \\ 2\zeta\omega_n &= T_2\omega_n^2 + 1/(T_1 + T_2) \end{aligned}$$

Solving for T_1:

$$\begin{aligned} T_1 &= \left[1/(2\zeta\omega_n - T_2\omega_n^2)\right] - T_2 \\ T_1 &= \left[1/\left\{1.414k - (707.2 \times 10^{-6})1{,}000k\right\}\right] - 707.2 \times 10^{-6} = 707.6 \times 10^{-6} \end{aligned}$$

For the AGC case, the constants are specified, and T_3 is solved as shown:

$$\begin{aligned} K &= K_aK_dK_c(1/T_3)(T_2/[T_1 + T_2]) \\ T_3 &= K_aK_dK_c(T_2/[T_1 + T_2])/K \\ K &= T_2\omega_n^2 \\ T_3 &= K_aK_dK_c(T_2/[T_1 + T_2])/\left(T_2\omega_n^2\right) \end{aligned}$$

For the PLL case, K_o and K_d are specified, and we solve for K_c:

$$\begin{aligned} K &= K_oK_dK_c(T_2/[T_1 + T_2]) \\ K_o &= 20K_d = 0.90 \\ K_c &= K/[K_oK_d(T_2/[T_1 + T_2])] \\ K &= T_2\omega_n^2 \\ K_c &= T_2\omega_n^2/[K_oK_d(T_2/[T_1 + T_2])] \\ &= 707.2 \times 10^{-6}(1.0 \times 10^6)/ \\ &\quad \left[20(0.9)(707.2 \times 10^{-6})/(707.6 \times 10^{-6} + 707.2 \times 10^{-6})\right] = 78.6 \end{aligned}$$

To complete the design, choose $C_1 = 0.01\ \mu\text{F}$:

$$T_1 = R_1C_1,\ \text{therefore}\ R_1 = T_1/C_1 = 707.6 \times 10^{-6}/0.01\ \mu\text{F} = 70.76\ \text{k}\Omega$$
$$T_2 = R_2C_1,\ \text{therefore}\ R_2 = T_2/C_1 = 707.2 \times 10^{-6}/0.01\ \mu\text{F} = 70.72\ \text{k}\Omega.$$

The resultant transfer functions are as follows:

$$^*T(\text{open loop}) = [707/S][S + 1.414k]/[S + 1.403k]$$
$$^*T(\text{closed loop}) = 707[S + 1.414k]/[S^2 + 2.11kS + 1{,}000k]$$

4.19 Similarities between the AGC and the PLL

The similarities between the PLL and the AGC are listed as follows. With very few exceptions, the analysis is nearly identical:

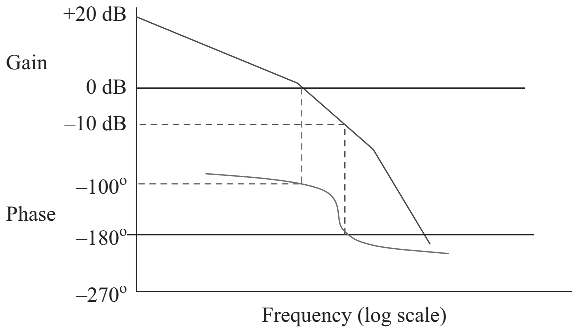
- The PLL has a built-in integrator, part of the VCO when connected into the loop.
- AGC has an additional time constant due to the added integrator $1/(R_3C)$.
- They have the same open-loop transfer function.
- They have the same closed-loop transfer function.
- They are basically the same process, since they are both feedback systems.

4.20 Feedback systems, oscillations, and stability

Feedback systems are valuable in reducing errors and maintaining constant values over changing parameters. Feedback helps the system evaluate the error from the ideal setting and allows the system to adjust values to minimize the error.

However, using feedback systems can cause the system to oscillate. If an oscillator is desired, then the design criteria are set to provide oscillations using feedback. If an oscillator is not desired, then the design criteria need to be implemented to avoid oscillations that are inherent in feedback systems. These types of systems use negative feedback, since positive feedback would cause the system to oscillate or continue to increase in value. Even with negative feedback systems, there is the possibility for system oscillation. For a system to oscillate, it has to follow the Barkhausen criteria, which is 0 dB gain and 360° phase shift. A negative feedback system, since it is negative, which by definition has a 180° phase shift, requires only a 180° phase shift for the system to oscillate: $180° + 180° = 360°$. If the negative feedback system has a 180° phase shift and 0 dB gain (gain of 1), then it satisfies the Barkhausen criteria and will oscillate.

Bode diagrams are used to provide an indication of the stability of a system and how much margin a system has from becoming unstable (Figure 4.15). This figure shows both the gain and phase as a function of frequency on a log scale. The gain margin is determined by the amount of additional gain required at a $-180°$ (negative feedback) phase shift to equal 0 dB gain, which is the oscillation criteria.



Instability & oscillation criteria for negative feedback systems, 0 dB Gain, –180° phase shift
Gain Margin (–180° phase shift) = 0 dB – (–10 dB) = 10 dB
Phase Margin (0 dB gain) = –100° – (–180°) = 80°

Figure 4.15 Bode plot for negative feedback systems

Since the gain is at -10 dB for a -180° phase shift, the additional gain required to equal 0 dB is:

$$\text{Gain margin} = 0\ \text{dB} - (-10\ \text{dB}) = 10\ \text{dB}$$

The phase margin is determined by the amount of additional phase shift required at 0 dB gain to equal a -180° (negative feedback) phase shift, which is the oscillation criteria. Since the phase shift is at -100° for a 0 dB gain, then the additional phase shift required to equal -180° is:

$$\text{Phase margin} = -100^\circ - (-180^\circ) = 80^\circ$$

4.21 Summary

AGC is an important element in the design of all types of transceivers. AGC provides the necessary DR required for operation over varying distances and conditions. AGC can be analyzed using control system theory because it is a feedback system. An integrator is required to provide a constant level out of the AGC to the detector providing zero steady-state error for an input amplitude step response. The similarities between AGC and PLL are remarkable. They both can be analyzed using modern control system techniques. The steady-state error for both systems is zero for a step response. The step response for AGC is related to a power level change, and the step response for PLL is related to a phase change. If a zero steady-state error is required for a step in frequency, then another integrator needs to be added. Further study of PLL is required to answer various questions, such as capture and lock range, operation of the loop in a no-lock situation, and considerations concerning stray poles that may change the simple analysis. However, the idea of using these control system techniques for analyzing both AGC and PLL can be useful.

4.22 Problems

1. Name at least two RF devices that can be used for gain control.
2. What would be a reasonable value of the RC time constant, following the diode detector for the AGC, for a carrier frequency of 10 MHz and a desired modulating frequency of 1 MHz?
3. What does an integrator in the feedback loop do for the steady-state error?
4. Explain how the integrator achieves the steady-state error in problem 3.
5. If the input signal to a receiver changes suddenly from −70 to −20 dBm, what is used to ensure a constant IF output of 0 dBm?
6. How does a nonlinear function affect the response of the AGC loop?
7. Why is a diode curve approximation much better than a piecewise linear approximation?
8. What is the main difference between an AGC and a PLL in the locked state?
9. What is the general reason that the AGC analysis is similar to the PLL analysis?
10. What state is the PLL assumed to be in the AGC/PLL analysis in this chapter?
11. What is the potential problem when using a feedback AGC?

Further reading

Bullock, S. R., and D. Ovard. “Simple Technique Yields Errorless AGC Systems.” *Microwaves and RF*, August 1989.

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Chapter 5
Demodulation

The demodulation process is part of the receiver process that takes the down-converted signal and retrieves or recovers the data information that was sent. This demodulation process requires three basic functions to retrieve the sent data:

- Recover the carrier, since the digital modulation results in a suppressed carrier and the carrier is recovered to remove it from the incoming signal
- Remove spread spectrum coding, if using spread spectrum techniques to mitigate jammers; generally done using a despreading matched filter correlator or a sliding correlator
- Align and synchronize the sample point for sampling the data stream at the optimal signal-to-noise ratio (SNR) point, which requires lining up the bits with the sample time using a bit synchronizer or over sampling the return.

The process detects the digital data that was sent from the transmitter with a minimum bit error rate (BER). This was performed in the past using analog circuitry, such as mixers and filters to remove the carrier frequency and spread spectrum codes, but today, the process is incorporated in the digital circuitry using application-specific integrated circuits, field-programmable gate arrays, and digital signal processing (DSP)-integrated circuits.

5.1 Carrier recovery for suppressed carrier removal

The carrier frequency is modulated in the transmitter by a digital waveform consisting of digital data or digital data combined with a high-speed digital pseudonoise (PN) code if spread spectrum is used. This process uses the digital waveform to shift the phase of the carrier according to the digital bit value or shift the frequency according to the bit values. In either case, the digital waveform creates a modulated signal that suppresses the carrier frequency, which is difficult to detect. For analog signals, basic techniques such as a phased-lock loop (PLL) can be used to determine the carrier frequency. A PLL uses a phase comparison between the carrier frequency and the local oscillator to recover the frequency. However, the PLL has a problem with fast-changing phases (digitally modulated carrier), which causes it to lose lock. Therefore, to recover the carrier, the following methods, known as carrier recovery loops, are used:

- Squaring loop
- Costas loop

- Modified Costas loop
- Decision-directed Costas Loop

Once the carrier has been recovered, it is used to cancel the suppressed carrier in the modulated waveform, which produces the digital waveform. These loops are discussed in detail in the following sections. There are trade-offs for which type of carrier recovery loop should be used. A careful look at these parameters will provide the best type of carrier recovery loop for use in any given system design.

5.1.1 Squaring loop

The squaring loop is employed for carrier recovery of a binary phase-shift keying (BPSK) direct sequence digital waveform. The advantage of the squaring loop is mainly its simplicity of implementation. The squaring loop requires minimal hardware, which makes it cost-efficient. A block diagram of a squaring loop is shown in Figure 5.1.

The squaring loop filters the incoming signal using a narrowband filter. The filter needs to be narrow enough for the spectrum of the input signal to be essentially constant. A filter matched to the waveform gives the best SNR and should be used in the intermediate frequency (IF) section before the squarer. This filter also reduces interference from other out-of-band signals and interference. The shape of the optimum filter is a $\sin(x)/x$ impulse response in the time domain (square wave function in the frequency domain). However, this filter is only theoretical and is not implementable. Therefore, for this application, a simple two-pole Butterworth filter can be used to approximate the ideal matched filter. The low-pass filter is designed to approximate the roll-off of the incoming waveform using a time period of T seconds, where T is one BPSK symbol interval. An alternative is to use an integrate-and-dump circuit, which also operates as a good matched filter.

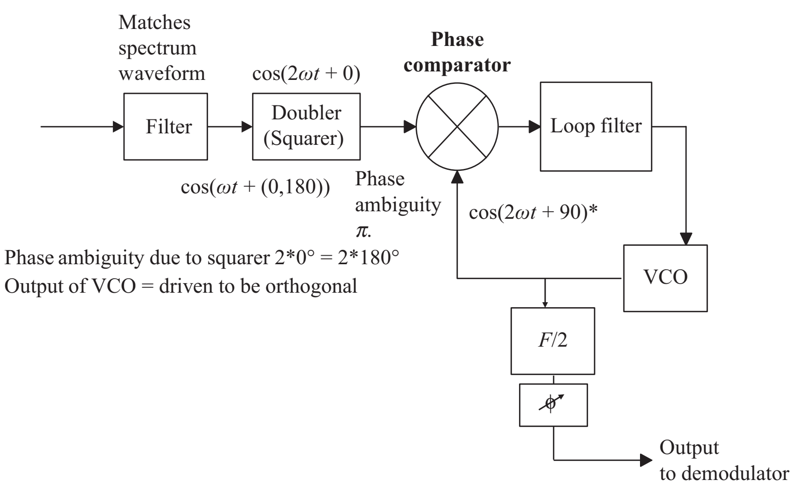


Figure 5.1 Squaring loop carrier recovery

The filtered signal is then fed into a doubler or squarer that basically squares the incoming signal. Squaring a direct sequence BPSK signal eliminates the modulation, since the modulated signal is phase shifted either 0° or 180°. Squaring the signal doubles both the frequency and phase, resulting an output of twice the frequency with no phase shift:

$$\begin{aligned} S &= A\cos(\omega t + \varphi), \text{where } \varphi = 0°, 180° \\ S^2 &= A^2(\cos^2(\omega t + \varphi)) = A^2/2(1 + \cos[2(\omega t + \varphi)]) \\ &= A^2/2 + A^2/2\cos(2\omega t + 2\varphi) \\ &= A^2/2 + A^2/2\cos(2\omega t + (0°, 360°)) \end{aligned}$$

Therefore,

$$S^2 = A^2/2 + A^2/2\cos(2\omega t + 0°)$$

The output is now an unsuppressed carrier at twice the frequency and thus can use a PLL to determine the frequency. Since the PLL detected frequency is at twice the desired frequency, the output frequency is divided by two to achieve the fundamental frequency and the carrier is recovered.

However, the divider creates an ambiguity since it divides whatever the phase is of the actual carrier by two. For example, a change in one cycle (2π radians) results in a change of phase of π. This means that if the frequency shifts one cycle, the phase is shifted 180° = 360°/2. Since the PLL does not recognize that there was a cycle slip, an ambiguity needs to be accounted for. This creates a potential problem in the data because of the possible phase reversal. Another example, suppose the frequency locks up at $\cos(2\omega t + (180°))$. When it is divided by two, the carrier frequency is $\cos(\omega t + (90°))$, which means that the carrier is 90° out of phase of the locked carrier. This phase needs to be adjusted out.

The output of the squaring loop is phase shifted to obtain a cosine wave, since the output of the PLL generates a sine wave and is multiplied with the incoming signal to eliminate the carrier and obtain the data. If the signal is 90° phase shifted with respect to the incoming signal, then the output would be zero:

$$\begin{aligned} (A\sin\omega t)(d(t)A\cos\omega t) &= A/2d(t)(\sin 0 + \sin 2\omega t) \\ &= A/2d(t)\sin 0 = 0, \text{ since } \sin 2\omega t \text{ is filtered} \end{aligned}$$

The phase shift is critical, may drift with temperature, and will need to be adjusted if different data rates are used. This is one of the main drawbacks to the squaring loop method of carrier recovery. Some other carrier recovery methods do not have this concern, though. The phase in question is the phase relationship between the input signal and the recovered carrier at the mixer where the carrier is stripped off.

Another disadvantage to the squaring loop approach to carrier recovery is that the filter for the squaring loop at the IF is good for only one data rate. If the data rate is changed, the filter needs to be changed. Some systems require that the data rate be variable, which makes the squaring loop less versatile than some other methods. A variable filter can be designed, but it is more complex than a fixed filter.

Squaring the signal gives rise to a direct current (DC) term and a tone at twice the carrier frequency. A possible disadvantage to the squaring loop is that it needs to operate at twice the frequency. For most cases this is not a problem, but higher frequencies can drive up the cost of hardware.

Another configuration of the squaring loop is shown in Figure 5.2. This configuration was developed for analog squaring loops, since it is generally easier to multiply frequencies than it is to divide frequencies in the analog domain. Since it is more difficult to divide frequencies in the analog domain, this configuration is another approach to achieve the same solution. Another advantage of this configuration is that the voltage-controlled oscillator (VCO) operates at a lower frequency, which is typically a cost savings. This configuration may not be an advantage in the digital domain since it is easier to divide using simple digital techniques than it is to multiply.

If higher order phase-shift keying (PSK) modulation schemes are used, then higher order detection schemes need to be implemented. For example, if a quadrature PSK QPSK modulation waveform is used, then the demodulation process should square the signal twice or raise the signal to the fourth power. Sometimes this circuit is referred to as a times 4 device, which is really a misnomer. The signal is not multiplied by 4, but the frequency is four times higher. Since with QPSK there are four phase states at 0°, 90°, 180°, and −90°, it requires a fourth power, or squaring the waveform twice, to eliminate the phase shift. The following example shows squaring the input two times with the amplitudes and DC terms ignored for simplicity:

$$\begin{aligned}(\cos(\omega t + (0°,90°,-90°,180°)))^2 &= \cos(2\omega t + (0°,180°,-180°,360°))\\ (\cos(2\omega t + (0°,180°,-180°,360°)))^2 &= \cos(4\omega t + (0°,360°,-360°,360°))\\ &= \cos(4\omega t + 0°)\end{aligned}$$

The frequency is recovered by dividing the resultant frequency by four.

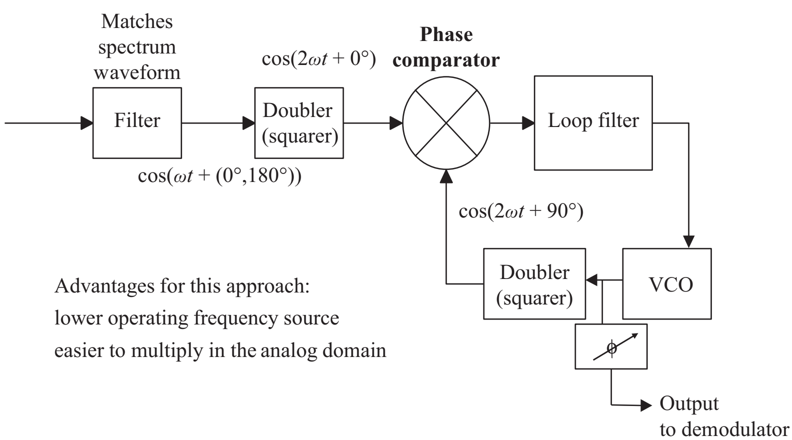


Figure 5.2 Another configuration of the squaring loop

The description of the squaring loop can be applied for all the higher order loops, keeping in mind that the functions need to be modified for the higher order operations. The difference is how many times the signal needs to be squared. For an n-phase signal, $2^a = n$, where a is the number of times the signal needs to be squared. For example, an 8-PSK signal would be $2^3 = 8$. Therefore, the 8-PSK signal would need to be squared three times to eliminate the phase ambiguity. The higher order PSK demodulation processes can become complex and expensive due to the higher frequency components that need to be handled. Also, there are losses in the signal amplitude every time it is squared. These losses are called squaring losses, and they degrade the ability to detect the signal.

Once the carrier is recovered, it is used to strip off the phase modulated carrier to retrieve the digital modulated signal. This is accomplished in the demodulator, or mixer, and is then integrated over a bit or chip time to achieve the digital signal. A large positive number from the multiply and integrate is a digital "1," and a large negative number from the multiply and integrate is a digital "0"; thus, the digital signal is retrieved (Figure 5.3). As noted before, this process can all be accomplished in the digital domain, which improves the reliability and performance.

5.1.2 Costas loop

The Costas loop is generally the preferred method of recovering the carrier frequency. It is used extensively in communication systems and does not require phase adjustment. It is basically a quadrature demodulation PLL and contains the complete process that eliminates the carrier frequency. This does not have the squaring loop phase ambiguity; however, it does have an ambiguity on which quadrature

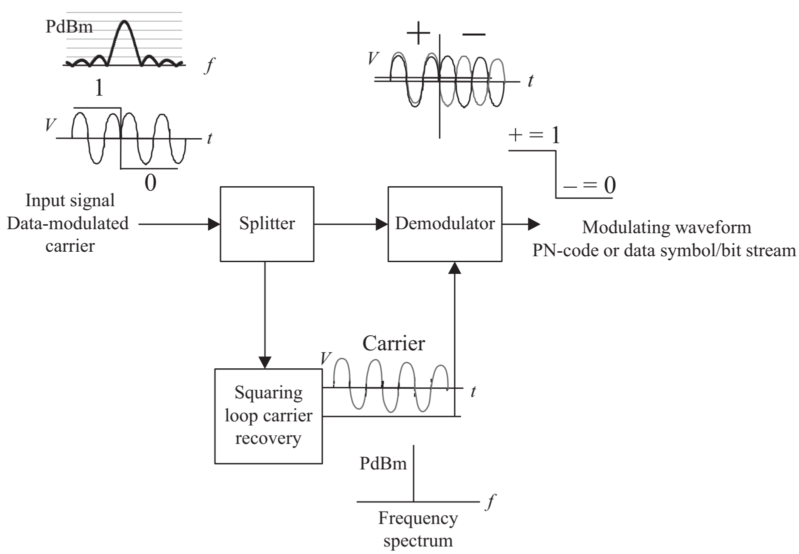


Figure 5.3 Digital signal recovery using a squaring loop

path contains the data, and in some systems the quadrature outputs are combined to mitigate this ambiguity. Another way to eliminate this ambiguity is to use a modified version of the Costas loop, known as a hard-limited Costas loop, which will be discussed shortly.

The standard Costas loop does not use a squaring loop, so the double-frequency signal is not a part of the process. Therefore, the higher frequencies are not produced. Instead, the Costas loop uses basically two PLLs, one of which is in-phase (I) and the other is in quadrature (Q) as shown in Figure 5.4.

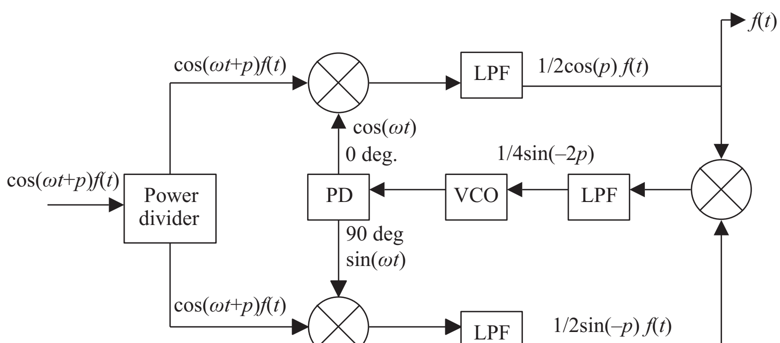
The outputs of each PLL are multiplied and low-pass filtered to eliminate the sum terms so that the loop tracks sin(2*P*), where *P* is the phase error. The low-pass filters are matched filters for the symbol or bit rate. To eliminate the need for an analog multiplier, a hard-limited Costas loop can be utilized. The matched filtering is done at baseband and the output is used to generate the error.

The phase shift, as in the squaring loop, is not critical, since the feedback forces the phase to be correct. Since the feedback is forcing the error to be zero, the error channel is driven to zero, but either channel can be the data channel. A modified Costas loop is a method to force the data to a known channel.

The matched filters in the Costas loop are good for only one data rate. If the data rate is changed, then the matched filters need to be changed. However, changing filters at baseband is much easier and more cost-effective than changing IF filters, as in the squaring loop. Note that the squaring loop baseband matched filters also need to be changed along with the IF filter.

5.1.3 Modified or hard-limited Costas loop and automatic frequency control addition

Since the Costas loop is good only for a narrow bandwidth, an automatic frequency control can be included to extend the pull-in range, which increases the bandwidth of the Costas loop. The standard Costas loop can be modified in other ways to improve the performance and versatility of this process. A hard-limited Costas loop



As *p* approaches zero the loop is locked. Note that the input could have been a sin instead of a cos and the bottom leg would have contained *f*(*t*).

Figure 5.4 Costas loop used for carrier recovery

or a data-aided loop can further improve the carrier recovery process. A hard-limited Costas loop has a hard limiter in one of the channels, which estimates the data pulse stream. By hard limiting one of the channels, that channel automatically, becomes the data channel since it produces a sign change, and the other channel becomes the error channel, which is driven to zero. The multiplier which is generally a drawback to Costas loops, since it is hard to build a multiplier in the digital domain can simply invert or noninvert the signal (Figure 5.5).

Using the data estimate, a noninverting amplifier or an inverting amplifier can be selected. This is much easier to implement in hardware. Commonly called a polarity loop, it strips the modulation off and leaves the phase error for the PLL. A data-aided loop uses the data estimation similar to the hard-limited Costas loop to improve the performance of the standard Costas loop. Many enhancements and variations can be made to the standard Costas loop, but a basic understanding of its operation provides the user with the ability to design a carrier recovery loop for a typical system, including modifications where improvement is desired.

5.1.4 Decision-directed Costas loop

The Decision-Directed Costas loop is used for quadrature signals including continuous phase-phase shift keying (CP-PSK). The basic diagram is shown in Figure 5.6. It is basically a quadrature hard-limited Costas loop where the decision

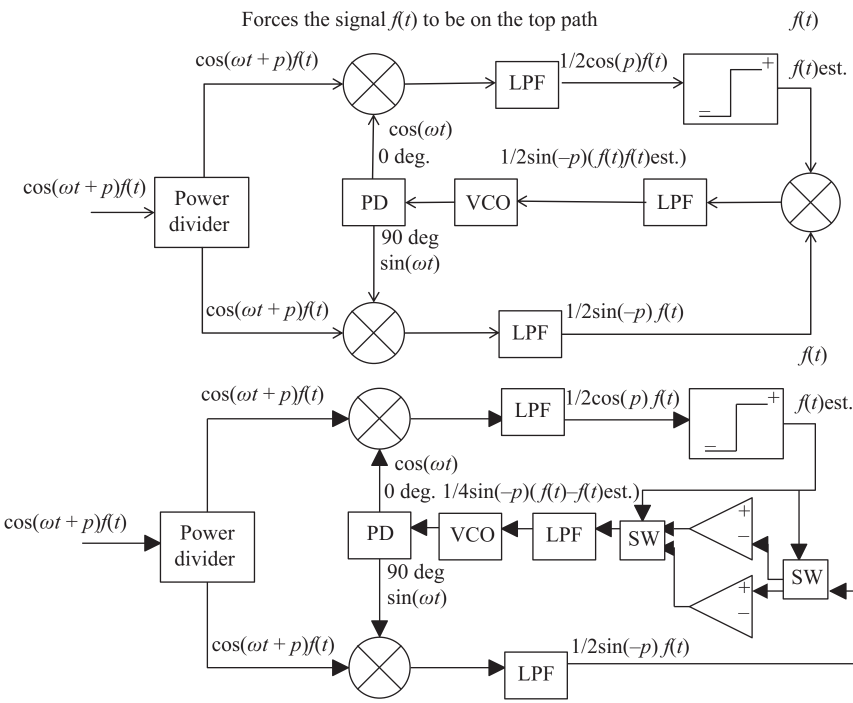
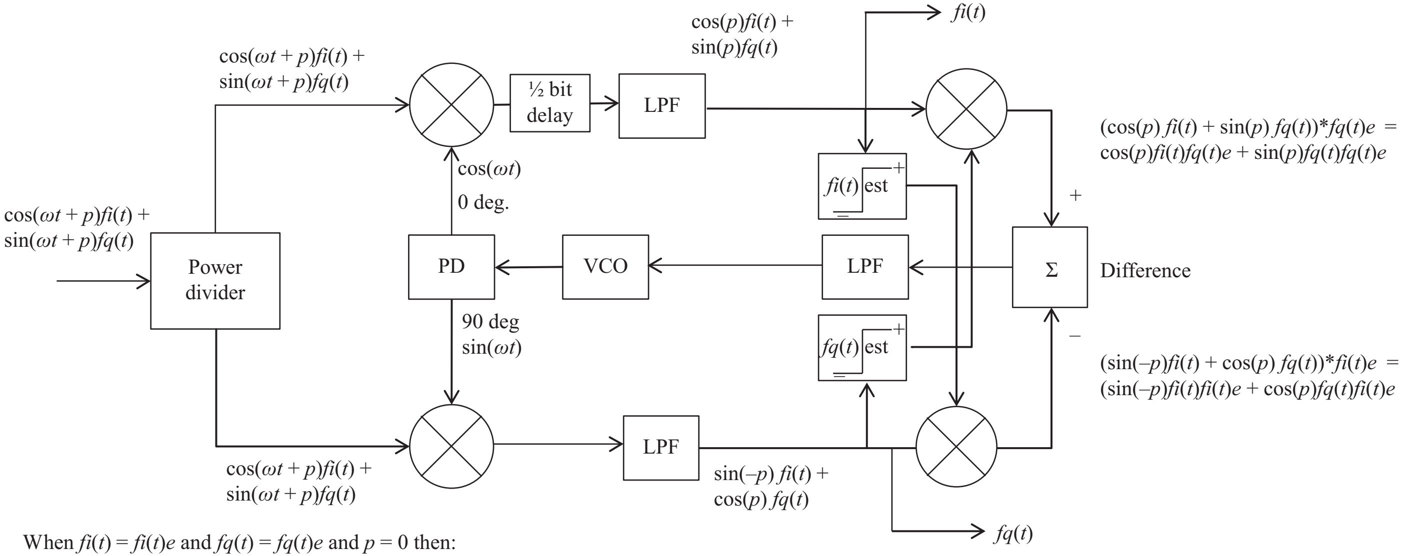


Figure 5.5 Modified or hard-limited Costas loop for carrier recovery for BPSK



When $fi(t) = fi(t)e$ and $fq(t) = fq(t)e$ and $p = 0$ then:
$\cos(0)fi(t)fq(t)$ for both outputs $= fi(t)fq(t)$. Subtracting these outputs = 0 and the PLL is locked.
The outputs are shown $fi(t)$ and $fq(t)$ after the LPFs.

Figure 5.6 Decision-directed Costas loop for quadrature signals including CP-PSK

is accomplished in both the I & Q channels, and both the channels are multiplied by the respected decisions and finally subtracted to produce the error for locking the VCO and Phase Lock Loop circuitry. Since the I & Q channels are offset by 1/2 bit, there is a 1/2 bit delay compensation in the I channel.

The I and Q channels of the differential CP-PSK waveform coming into the decision directed Costas Loop is shown in Figure 5.7. Since it is differential, a "0°" shows no change in phase and a "1" shows a 180° change in phase for both the I and Q channels independently. Note that the Q channel is offset 1/2 of a bit time so that they do not change simultaneously. The sum of both the I and Q channels provides the decoded data bits for the output. Both the I & Q channels are sinusoidal and summing them together produce a continuous wave (CW) in the time domain for the CP-PSK waveform (Figure 5.8). This is basically the same waveform that is produced in the transmitter in ideal conditions.

5.2 Demodulation process to remove spread spectrum code

Once the carrier is removed, the remaining digital waveform is a combination of the spread spectrum PN code and the digital data. To retrieve the digital data, the PN code needs to be removed or stripped off of the digital waveform. There are basically two ways to demodulate this digital waveform for PSK signals: a matched filter correlator or a sliding correlator. The method used depends on the type of modulation waveform and the corresponding demodulation process.

5.2.1 Matched filter correlator

Another option to remove the spread spectrum PN code is to use an asynchronous process that combines the phase-shift times (bits or chips) using a matched filter correlator. This matched filter correlator generally consists of a tapped delay line, with the tap spacing equal to the bit or chip duration, and provides a means to sum all of the phase shifts. This can be done either using analog tapped delay lines or digital filters. Most systems use digital means for implementing this type of demodulation scheme.

As the digital signal propagates through the matched filter, a correlation peak is generated at a particular time. The peak can be used to retrieved data directly or in a pulse position modulation (PPM) scheme where the time position of the correlation peak decodes the data.

The matched filter process is used for asynchronous detection of a spread spectrum signal. It can be used for pulsed systems where the arrival of the pulse provides the information for the system. The time of arrival (TOA) is used in conjunction with a PPM scheme to decode the sent data. This is very useful for systems that do not transmit continuous data and that do not want to use extra bits (overhead bits) for synchronizing the tracking loops in a coherent system. This type of matched filter is a very simple process that sums the PN signals in such a way that the signal is enhanced, but it does not change the bandwidth.

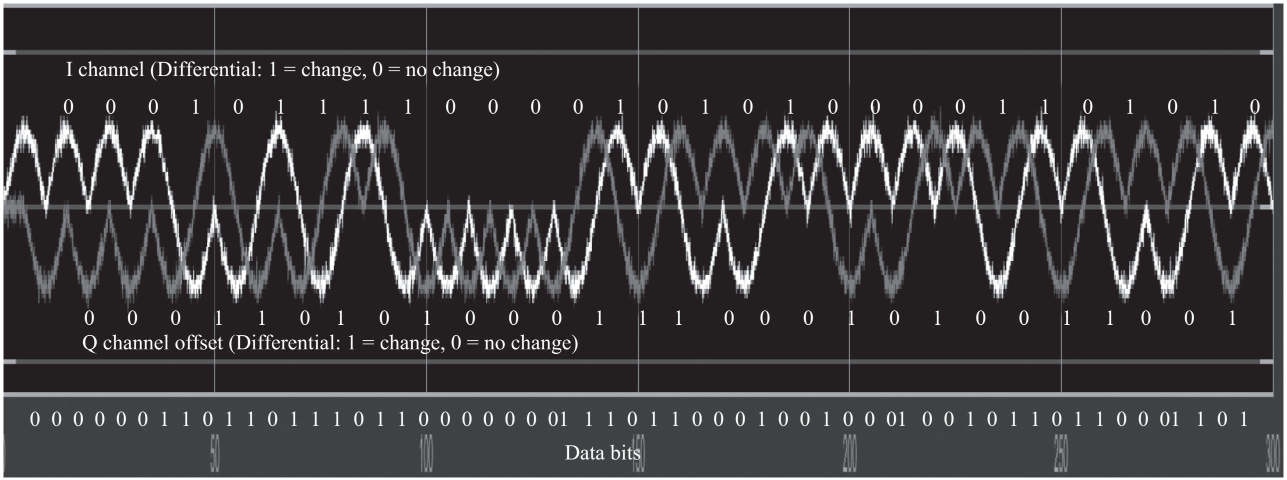


Figure 5.7 Differential CP-PSK I & Q channels out of the decision directed Costas loop

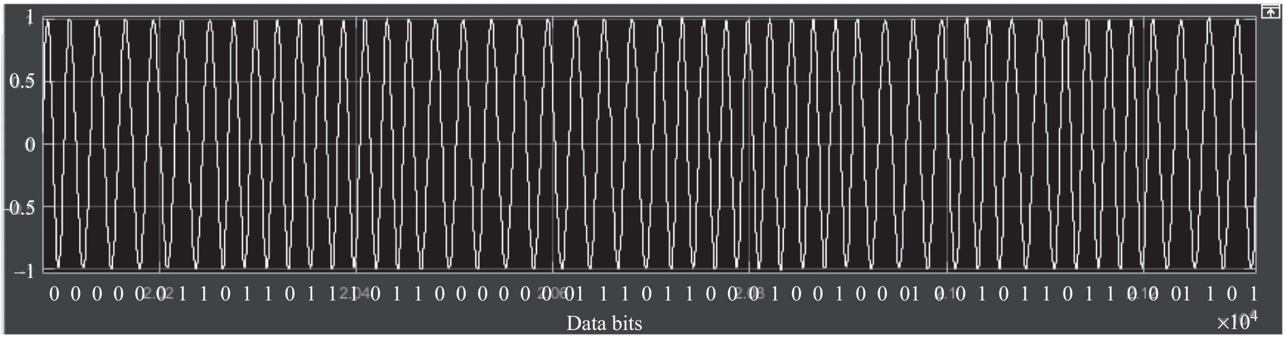


Figure 5.8 Differential CP-PSK by combining the I & Q channels

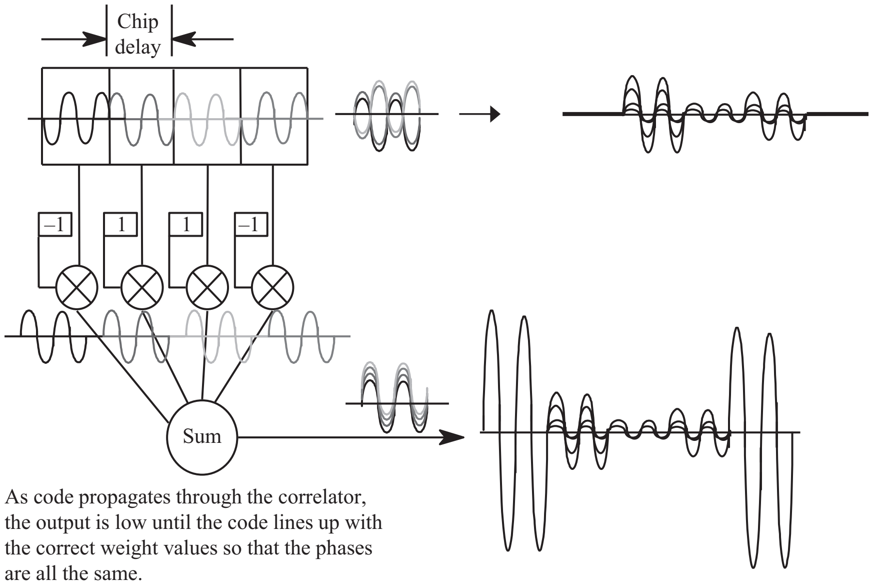


Figure 5.9 Analog matched filter correlator

An example of this type of matched filter correlator in the analog domain is an acoustic charge transport device. It separates packets of sampled analog data with respect to delay, weights each packet, and then sums them together to receive a desired pulse. An analog version of a matched filter is shown in Figure 5.9. The analog method is generally not used due to the problems with delay changes and phase shifts over temperature and other phase distortion that occurs in the analog domain.

The matched filter is very useful in the digital domain, and this process can be done digitally using finite impulse response filters with the tap delay equal to the chip width and weights that are 1 or -1, depending on the code (Figure 5.10). The matched filter correlator provides a means of producing a TOA with a high SNR by combining all of the pulses that were sent into one time frame.

The matched filter correlator consists of a tapped delay line, with the delay for each of the taps equal to 1/chip rate. Each of the delay outputs are multiplied by a coefficient, usually ± 1, depending on the code that was sent. The coefficients are the time reverse of the PN code values. For example,

Code values: 1, -1, 1, 1
Coefficient values: 1, 1, -1, 1

As the signal is processed through the matched filter, the weights will correspond at one code delay in time, and the outputs of the resultants are all summed together to create a large signal (Figure 5.10).

This example shows a code that is 4 bits long. The output of the digital summation is a pulse with an amplitude of four times the input pulse amplitude.

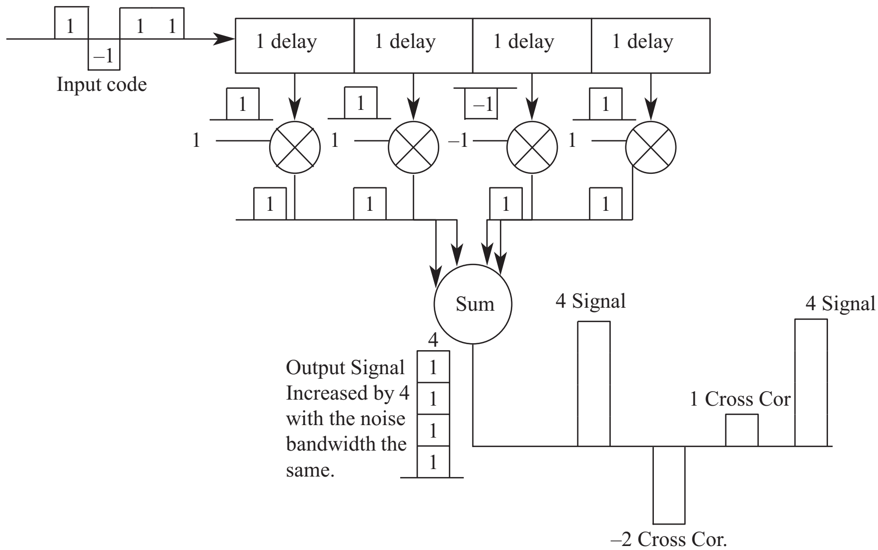


Figure 5.10 PN-code digital matched filter correlator for spread spectrum systems

This provides the SNR necessary to detect the signal with the bandwidth and noise power unchanged.

Often digital systems will use a quadrature down conversion to strip off the carrier and provide both I and Q data to the matched filter. The matched filter can employ either a quadrature correlator or parallel matched filters to process the I and Q data.

To increase the process gain and jamming margin, cascaded PN-code-matched filters can be used. The more cascading of the match filters, the more process gain and jamming margin is achieved. An example of cascaded matched filters is shown in Figure 5.11. This simple example shows a four-stage matched filter, achieving an increase in signal of four times, then the output of this match filter is fed into another four-stage matched filter to provide an increase in signal level of 16 times (Figure 5.11).

If the code is longer and random, then passing a sine wave through it would result in superposition of many noninverted and inverted sine wave segments. When these segments are summed together in the matched filter, the average output would be close to zero amplitude, or at least reduced. The same thing applies with different codes that are passed through the matched filter correlator, since they switch the phase of the carrier at the wrong places and times. The closer the code is to the pseudorandom code, the higher the output correlation peak will be. Note, however, that regardless of the code used, the bandwidth has not changed. The bandwidth is dependent on the chip rate or pulse width (PW), which is still two times the chip rate (or 2/PW) for double-sided, null-to-null bandwidth (Figure 5.12).

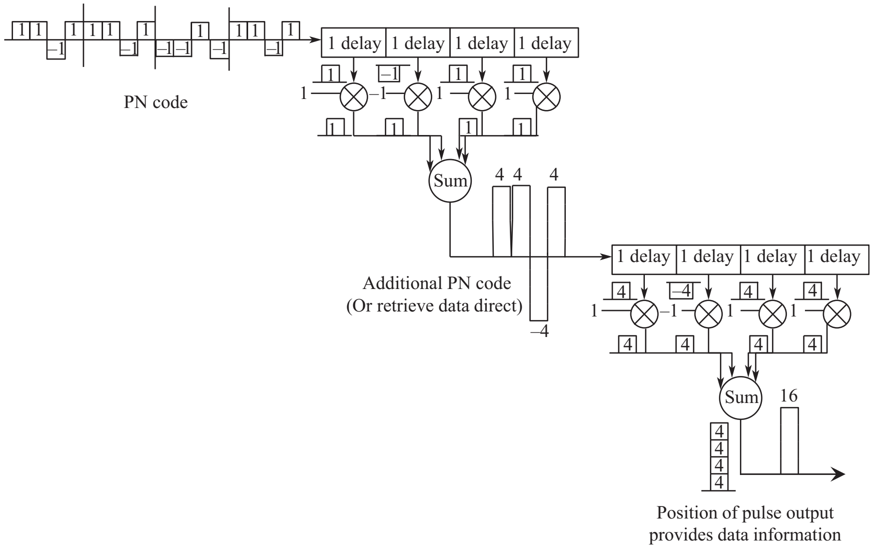


Figure 5.11 Cascaded PN-code-matched filters for increased process gain and margin

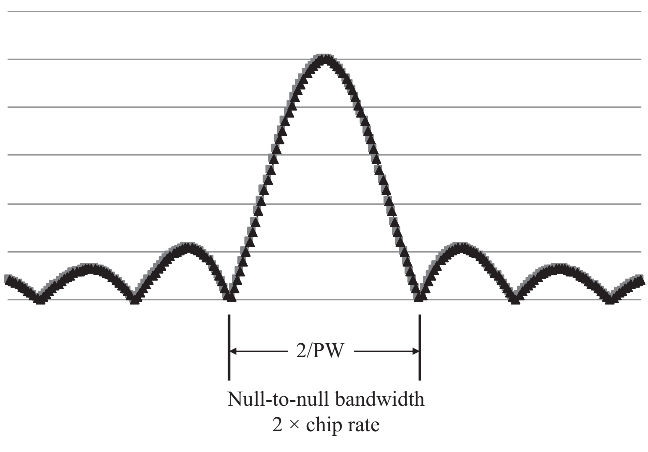


Figure 5.12 Sinc function squared for pulse modulation

Frequently, there is no carrier recovery since the process is asynchronous and the signal is demodulated. This method of detection is useful in pulsed systems or time division multiple access (TDMA) systems, where the overhead required to synchronize the system may overwhelm the amount of data to be sent. In other words, the overhead bits and time to synchronize the tracking loops reduces the amount of information that can be sent for a given transmission PW.

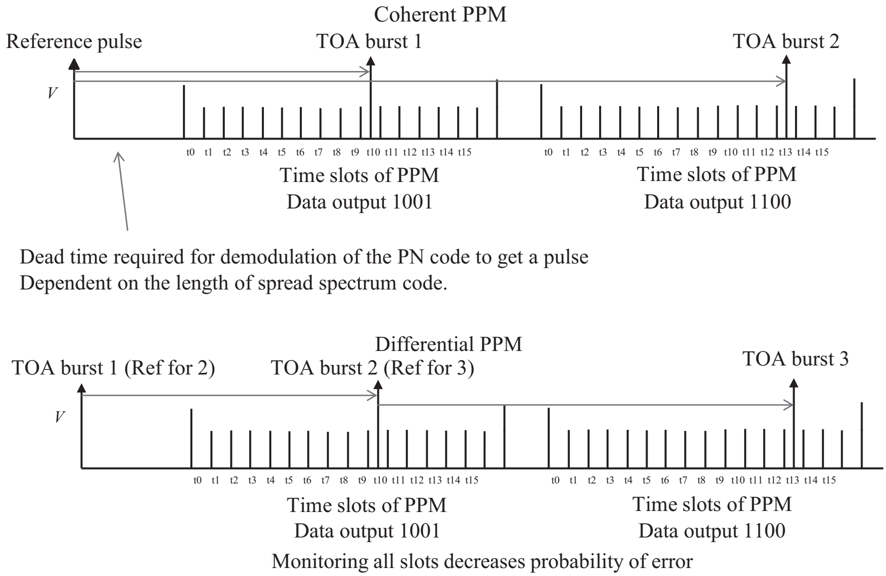


Figure 5.13 Coherent and differential pulse position modulation (PPM) for data encoding

Once the pulse is recovered, this produces a pulse in time that establishes the TOA for that pulse. This can be applied to a PPM technique to demodulate the sent data.

5.2.1.1 Pulse position modulation

PPM measures the TOA of the pulse received to determine where in time the pulse occurred. The data are encoded and decoded with reference to the time position of the pulse. The data bits represent a time slot on the time slot grid array. This is used with spread spectrum systems using PN codes, with matched filters to correlate PN codes. The integrated correlated signal produces a pulse, and the time position of the pulse provides data information.

The number of bits is dependent on the number of time slots; for example,

$$2^n \text{ bits} = \text{number of time slots}$$

If there are eight time slots, then there are 3 bits of information for each time slot:

$$2^3 \text{ bits} = 8 \text{ time slots}$$

If absolute time is precisely known, then an absolute PPM demodulation process can be used. Therefore, the TOA referenced to absolute time provides the information. The amount of information this TOA pulse provides depends on the PPM grid or the number of possible time divisions in which the TOA could occur. For example, if there are eight possibilities for the TOA to occur, then 3 bits of information are produced on the TOA of one pulse (Figure 5.13).

The PPM grid is referenced to an absolute time mark, and the entire PPM grid changes only with absolute time variations. So the PPM grid is not dependent on the TOA pulses received except for the reference. Once the PPM structure is set up, it does not change. The dead time is allocated to ensure that the chips that make up the pulsed output have passed through the matched filter and do not interfere with other pulses. The dead times are the same lengths and do not vary in length between pulses.

Differential PPM also can be useful when absolute time is not known. This method relies on the difference in time between two received TOA pulses. Therefore, the first TOA pulse received contains no data but provides the reference for the following TOA pulse received. The time after a received TOA pulse is divided up into a PPM time slot grid, and when the following TOA pulse is received in that time slot grid then the time slot determines the sent data. For example, if the time on the grid is divided into eight time slots after receipt of a TOA pulse, then the next TOA pulse falls in one of the eight time slots and produces 3 bits of information (Figure 5.13). The third TOA pulse is mapped into the PPM grid, with the second TOA pulse taken as the reference, and so on, as shown in Figure 5.13.

Dead time is also required for this PPM scheme. However, the dead time spacing between the PPM grids is contingent on the location of the received reference pulse. For example, if a TOA pulse is received in time slot 1 of the PPM grid, it is used as a reference for the next TOA pulse. The dead time starts from the time of reception of the reference TOA pulse. If the same TOA pulse is received in time slot 8 of the PPM grid and is used as the reference for the next pulse, the absolute time of the PPM grid for the next pulse will be delayed more than if time slot 1 was used as the reference. The average would produce a higher throughput of data compared with absolute PPM, since the dead time starts immediately following a received TOA pulse. For the absolute time PPM, the grid for all pulses is at the maximum and does not change with the data used.

For many applications, it is better to leave the grid constant so the entire grid is received to determine the highest target in case there are jammers or large signals from interference.

A comparison between absolute and differential PPM with respect to missed pulses is shown in Figure 5.14. With absolute PPM, if a pulse is missed or appears in another time slot, then the errors will be for that time slot only. However, if differential PPM is used, then a TOA pulse that is missed or falls into the wrong time slot would cause an error not only for that particular time slot but also for the next TOA pulse position because the time slot grid is establish by the previous TOA pulse that was in error (Figure 5.14). Therefore, differential systems have a possibility of having twice as many errors compared to the absolute system. This is true for all types of differential systems and will be discussed in other applications.

5.2.2 Sliding correlator

A sliding correlator is used to remove the spread spectrum code on a spread spectrum system. This process incorporates a method of stripping off the PN code

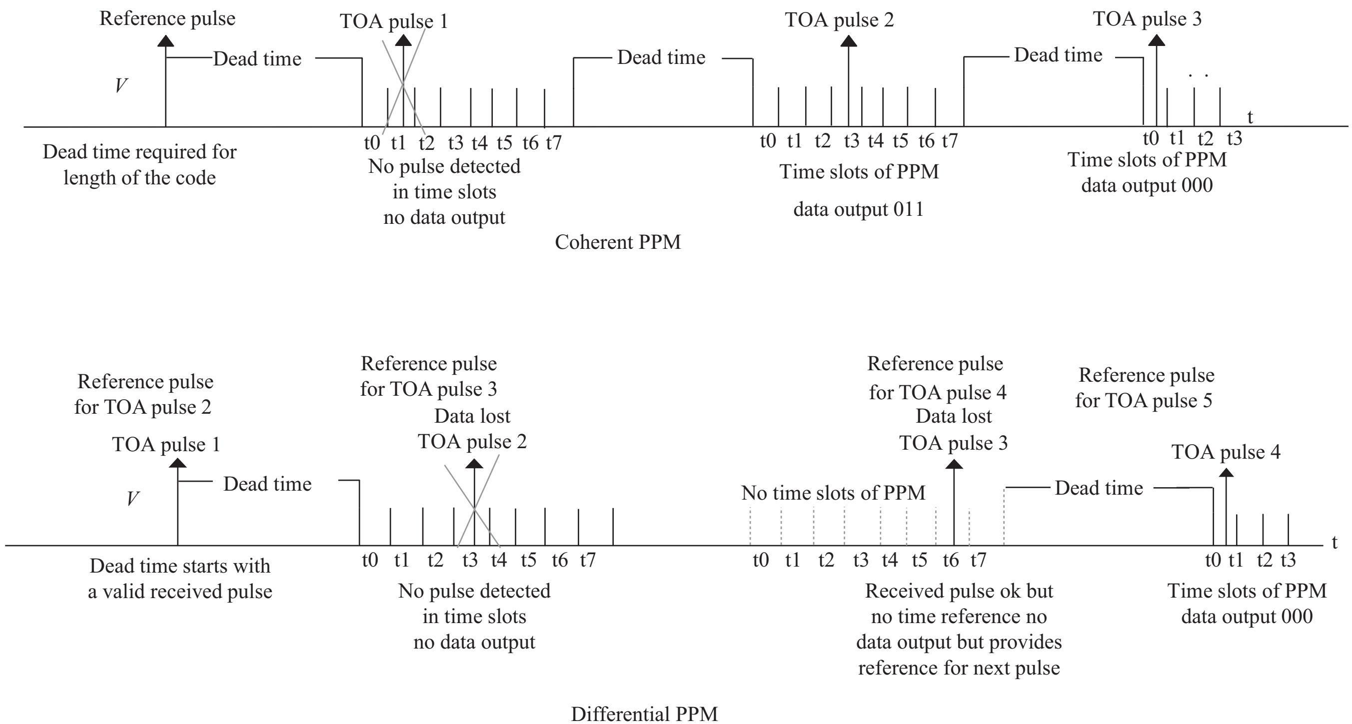


Figure 5.14 Comparison between coherent and differential PPM

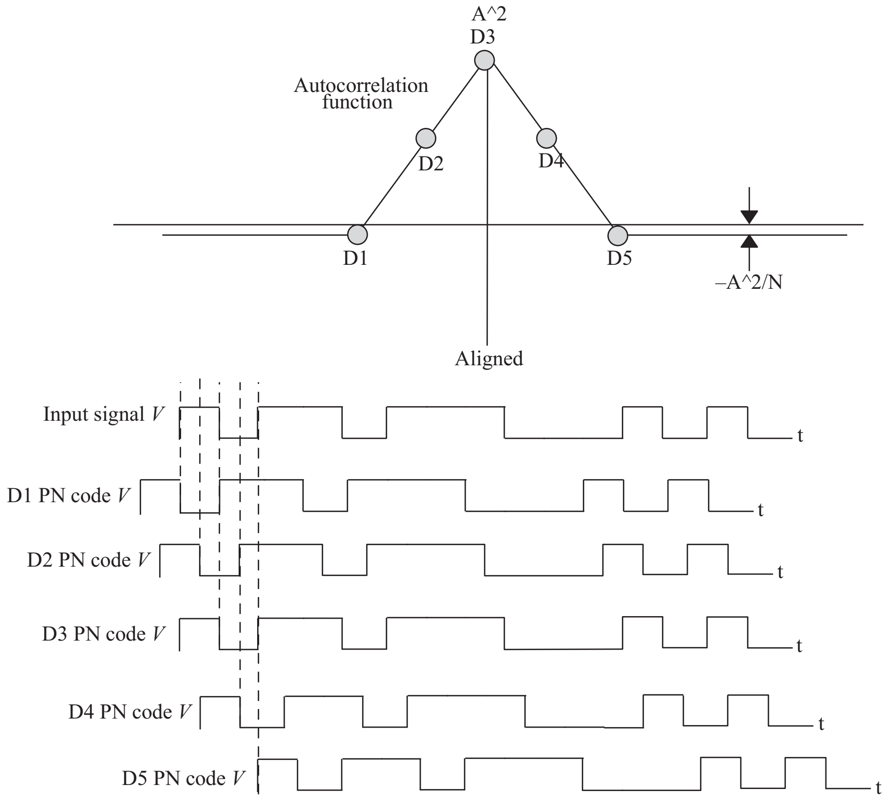


Figure 5.15 Sliding correlator strips off the PN-code

to retrieve the data information. The receiver has the same PN code generator and code as the transmitter so the technique is to align the incoming code with the received code. The correlator slides the code in time with the incoming code and looks for a correlation peak (Figure 5.15). D1 and D5 are not lined up and do not have a correlation value. D2 and D4 have a 1/2 correlation value, and D3 has a full correlation value or the peak of the correlation. This is where the sliding correlator has found the code timing and can strip off the code. The total correlation of the entire code approaches zero and can be stripped off leaving only the data.

When the peak is found, the system switches into lock condition and tracks the changes to keep the codes aligned. Once they are aligned, the codes are multiplied and integrated, which removes the PN code and retrieves the digital data. Since the code can be long before it repeats, some codes do not repeat for 2 years so an initial alignment needs to be performed. This requires additional bits to be sent that are not related to the data being sent. These bits are called the preamble, which consists of overhead bits that will lower the data rate by a certain percentage of the data messages sent. For continuous systems, these overhead bits are a very small percentage of the message. For pulsed-type systems or TDMA systems, these bits can

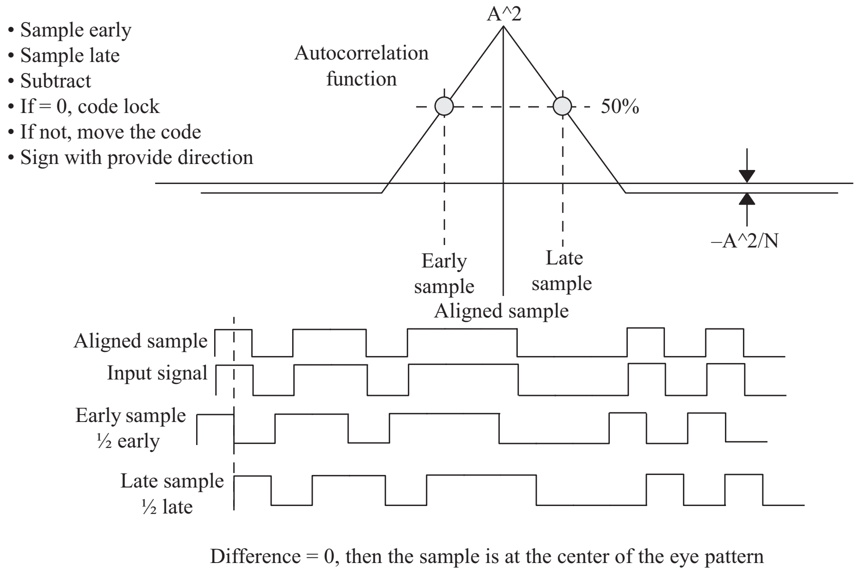


Figure 5.16 Correlation peak for early-late gate alignment

be a large percentage of the message depending on the length of the pulse. The code alignment process can be accomplished by using the following techniques:

- Generating a short code for acquisition only
- Using highly reliable clocks available to both the transmitter and receiver, such as rubidium, cesium, or global positioning system (GPS)
- Using an auxiliary subsystem that distributes system time to both parties.

In many system architectures, all of these techniques are used to simplify and accelerate the code alignment process. Once the code time is known fairly accurately, two steps are used to demodulate the PN code.

The first step is to determine if the code is lined up; the second step is to lock the code into position. The first step is achieved by switching a noncorrelated code into the loop and then into a sample-and-hold function to measure the correlation level (noise) output. Then the desired code is switched into the loop and into the sample-and-hold function to measure the correlation level (signal) output. The difference in these levels is determined, and the results are then applied to a threshold detector in case the search needs to be continued. If the threshold level is achieved, the second step is activated. This prevents false locking of the sidelobes. Note that a false lock generally occurs at half of the bit rate.

The second step is to maintain lock (also fine-tuning) by using a code lock tracking loop. This assumes that the codes are within 1/2 bit time. One type of code tracking loop is called an early-late gate. An early-late gate switches a VCO between an early code (1/2 bit time early) and a late code (1/2 bit time later). The correlation or multiplication of the early code and the input code and the late code and the input code produces points on the autocorrelation peak (Figure 5.16).

The code loop is locked when the values of the autocorrelation function are equal. The nondelayed code is used for demodulation of the spread spectrum waveform. The peak of the autocorrelation function occurs when the nondelayed aligned code is multiplied with the incoming signal and summed (Figure 5.16). These codes are mixed with the incoming signal, filtered, and detected (square-law envelope detected), and the output sign is changed at the dither rate. For example, the early code detector output is multiplied by -1, the late code detector output is multiplied by $+1$, and the final output is filtered by the loop filter, which takes the average of the two levels. This provides the error and controls the VCO to line up the codes so that the error is zero, which means the VCO is switching symmetrically around the correlation peak. Therefore, the early-late gate uses feedback to keep the code aligned. Sign is important in the design and hardware applications since it drives the direction of the oscillator. For example, if a negative number increases the frequency of the VCO or a positive number increases the VCO, this needs to be taken into account so the VCO is driven in the right direction given the sign of the error signal.

A delay-locked loop (DLL) is another form of a lock loop code. DLLs accomplish the same thing as early-late gates. They align the code to strip off the PN code using autocorrelation techniques. The main difference in the DLL compared with the early-late gate is that the DLL uses a nondelayed code and a code that is delayed by one chip. The DLL splits the incoming signal and mixes with the VCO output that is either shifted a bit or not. The main goal is to obtain the maximum correlated signal on the nondelayed code and the minimum correlated signal on the 1 chip delayed code. The DLL generally tracks more accurately (approximately 1 dB), but it is more complex to implement. Other forms of correlation that are used in communication systems include tau-dither loops and narrow correlators. Narrow correlators are used in GPS systems for higher accuracy and will be discussed later.

A basic code recovery loop is shown in Figure 5.17. The incoming code is a composite of both the data and the high-speed PN code from the transmitter to

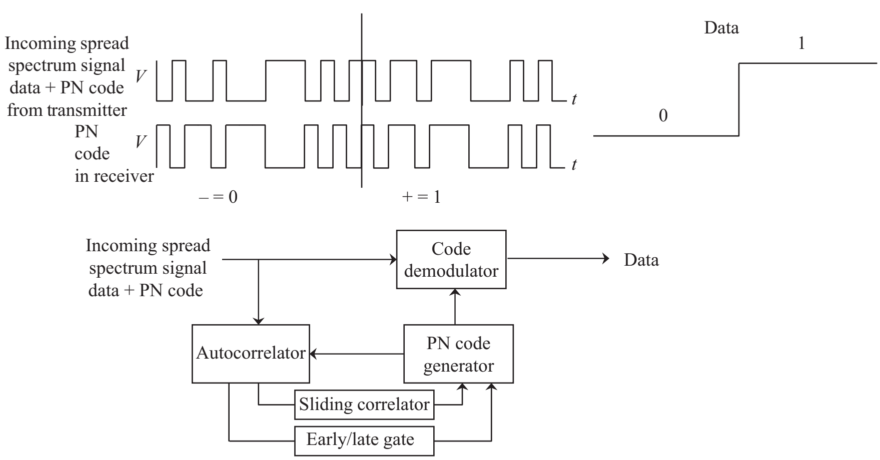


Figure 5.17 Code recovery example

produce the spread spectrum signal. The same high-speed PN code resides in the receiver. The sliding correlator searches for the correlation peak by speeding up or slowing down the code clock, which aligns the codes. When these codes are aligned, multiplication is performed on each of the code bits and the integration of all the multiplication products results in a large negative number when the code is inverted, which represents a “0” data value in this example; the same process results in a large positive number when the code is not inverted, which represents a “1” data value. Thus, if the codes are aligned properly, the output is the retrieved digital data signal (Figure 5.17). The autocorrelator, or early-late gate in this example, compares the incoming digital waveform and correlates this digital signal with the internal PN code residing in the receiver. Using an early code and a late code or early-late gate and comparing these results provides the feedback to adjust the speed of the PN receiver clock to maintain code lock.

5.3 The eye pattern

The eye pattern describes the received digital data stream when observed on an oscilloscope. Due to the bandwidth limitations of the receiver, the received bit stream is filtered, and the transitions are smoothed. Since the data stream is pseudo-random, the oscilloscope shows both positive and negative transitions, thus forming a waveform that resembles an eye (Figure 5.18).

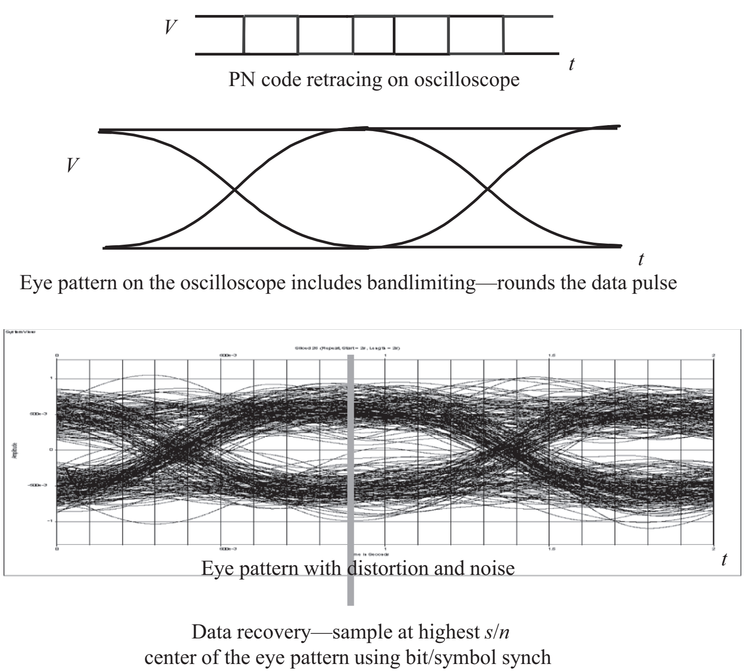


Figure 5.18 Eye pattern as seen on an oscilloscope

The four possible transitions at the corner of the eye are low to high, high to low, high to high, and low to low. Observation of the eye pattern can provide a means of determining the performance of a receiver. The noise on the eye pattern and the closing of the eye can indicate that the receiver's performance needs to be improved or that the signal from the transmitter needs to be increased. The eye pattern starts to close with the amount of distortion and noise on the signal (Figure 5.18). The bit synchronizer samples the eye pattern at the center of the eye where the largest amplitude and the highest SNR occur. Over sampling can also be used to determine the highest SNR. As the noise increases, it becomes much more difficult to determine the bit value without bit errors (Figure 5.18).

5.4 Intersymbol interference

Intersymbol interference (ISI) is the amount of interference due to dispersion of the pulses that interfere with the other pulses in the stream. Dispersion occurs when there are nonlinear phase responses to different frequencies or a nonconstant group delay, which is the derivative of the phase (Figure 5.19).

Since pulses are made up of multiple frequencies according to the Fourier series expansion of the waveform, these frequencies need to have the same delay through the system to preserve the pulse waveform. Therefore, if the system has a constant group delay, or the same delay for all frequencies, then the pulse is preserved. If a nonconstant group delay is causing different delays for different frequencies, then these frequencies are added together to form a pulse waveform where the pulse is dispersed or spread out. This leads to distortion to adjacent pulses and thus creates ISI. To obtain a value of the amount of ISI, the eye pattern can be used. The ISI is determined by the following and is shown in Figure 5.19.

$$\text{ISI} = 20\log(V_h/V_l)$$

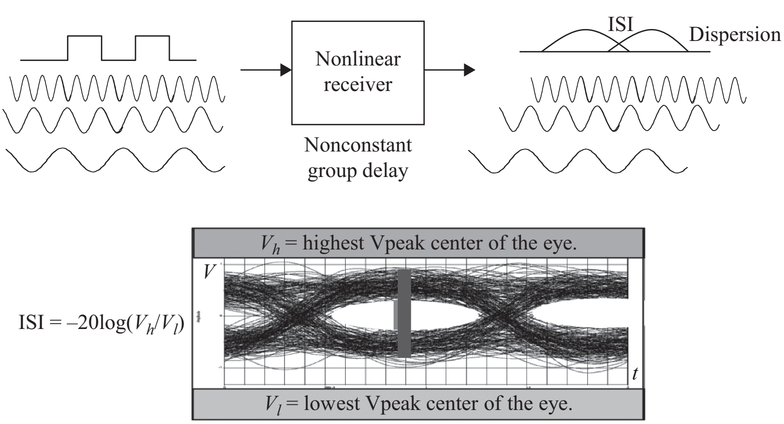


Figure 5.19 Intersymbol interference as a result of dispersion of the pulse waveform

where V_h is the highest peak voltage when measured at the center of the eye of the eye pattern and V_l is the lowest peak voltage when measured at the center of the eye of the eye pattern.

The measured amount of ISI in a receiver determines the amount of degradation in the ability to detect the desired signals and leads to increased BER in the system. This can be included in the implementation losses of the system, see Chapter 1. Sometimes, this ISI is specified in the link budget as a separate entry.

5.5 Symbol synchronizer

Once the carrier is eliminated and the spread spectrum PN code is removed, the raw data remain. However, due to noise and ISI distorting the data stream, a symbol synchronizer is needed to determine what bit was sent. This device aligns the sample clock with the data stream so that it samples in the center of each bit in the data stream. An early-late gate can be used, which is similar to the early-late gate used in the code demodulation process. The latter requires integration over the code length or repetition to generate the autocorrelation function, but with the bit synchronizer, the integration process is over a symbol or bit if coding is not used. The early and late gate streams in the bit synchronizer are a clock of ones and zeros since there is no code reference and the data are unknown. Therefore, the incoming data stream is multiplied by the bit clock, set at the bit rate, with the transitions early or later than the data transitions, as shown in Figure 5.20.

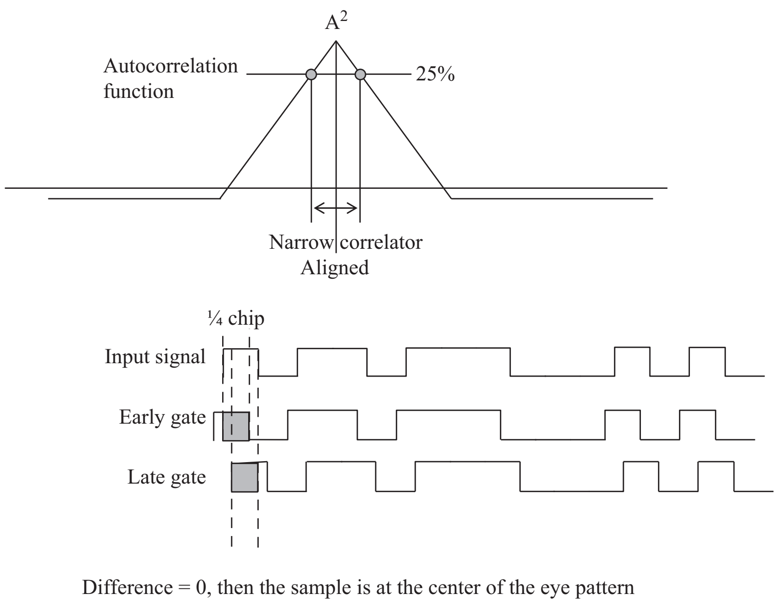


Figure 5.20 Symbol synchronizer using early-late sampling with a narrow correlator

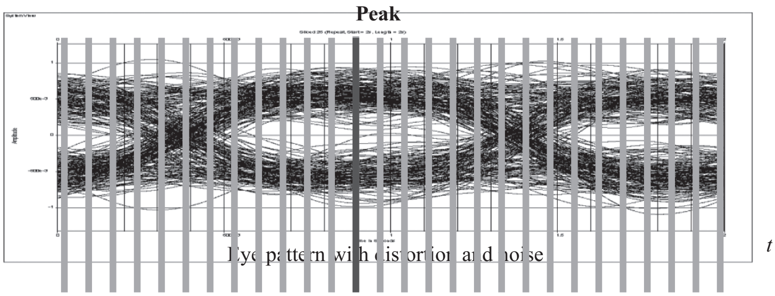


Figure 5.21 Determine peak amplitude using multiple samples

This example shows the early gates and late gates are off from the aligned signal by 1/4 symbol or bit, which is a narrower correlator compared with the standard 1/2. This provides more accuracy for the sample point. The integrated outputs are shown along with the aligned integration for reference. When the integrated outputs are equal in peak amplitude, as shown in the example, then the bit synchronizer is aligned with the bit. So, the point between the early gate and late gate, which is the center of the aligned pulse, is the optimal point to sample and recover the data. This point provides the best SNR of the received data, which is the center of the eye pattern (Figure 5.20). Once these data are sampled at the optimal place, a decision is made to determine whether a "1" or a "0" was sent. This bit stream of measured data is decoded to produce the desired data.

5.6 Digital processor

An alternative to the symbol synchronizer uses multiple samples of the eye pattern to determine the peak from the results of the samples (Figure 5.21). The levels of each of the samples is measured and processed to determine the maximum point for the decision point. This is generally the most common approach to determine the data values from the eye pattern and does not require the early-late gate processing.

Once the digital data stream has been sampled and it is determined that a "0" or a "1" was sent, the digital data stream is decoded into the data message. The digital processor is generally responsible for this task and also the task of controlling the processes in the receiver as well as demodulation of the received signal. In many systems today, the digital processor, in conjunction with specialized DSP integrated circuits, is playing more of a key role in the demodulation process. This includes implementation of the various tracking loops, bit synchronizing, carrier recovery, and decoding of the data, as mentioned already. The digital processor ensures that all of the functions occur at the necessary times by providing a control and scheduling capability for the receiver processes.

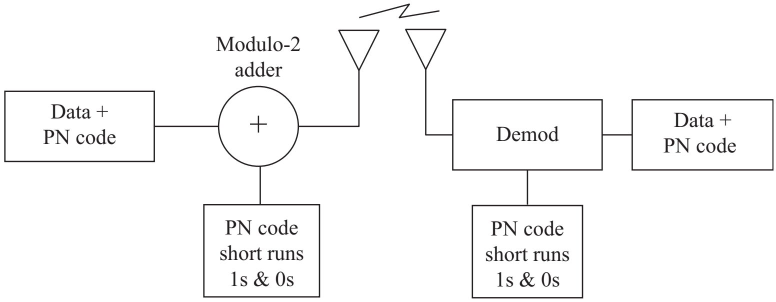


Figure 5.22 Scrambler/descrambler prevents long streams of "1"s and "0"s

5.7 Scrambler/descrambler

Scramblers and descramblers are used to prevent a long stream of either "1"s or "0"s in the data or code. The scrambler uses an additional PN code selected with short runs of "1"s and "0"s which is combined with the code or data sequence. This is accomplished by performing a modulo-2 addition to the digital signal in the transmitter (Figure 5.22). This code is demodulated by the descrambler in the receiver with a matched filter correlator or sliding correlator. The code used for the scrambler/descrambler contains no data and adds cost and complexity. Using this technique prevents drifts of VCOs due to large DC values caused by these long periods when the code does not change state.

5.8 Shannon's limit

In 1948, C.E. Shannon developed a theory of how much information could be theoretically sent over a channel. The channel could be any medium used to send information, such as wire, optic cable, or free space. The Shannon limit for information capacity is:

$$C = B\log_2(1 + \mathrm{SNR})$$

where C is the information capacity of a channel (bps), B is the bandwidth (Hz), and SNR is the signal-to-noise ratio.

Note: *SNR* is the actual value of the SNR, which is unitless and is not converted to dB. So if the SNR is given in dB, the value needs to be reverted to the actual SNR.

Since it is more common to use logarithms in base 10 instead of base 2, a conversion is required as follows:

$$\log_2 X = \log_{10} X/\log_{10} 2 = 3.32\log_{10} X$$

Therefore, Shannon's limit becomes:

$$C = 3.32B\log_{10}(1 + \mathrm{SNR})$$

Many different coding schemes can send several bits of information for every symbol sent. For example, QPSK sends 2 bits of information per symbol of transmission. Shannon's limit provides the absolute maximum information that can be sent no matter how many bits can be put into a symbol. This is defined for a single channel, so using multiple parallel channels such as MIMO to increase the data rate does not disprove Shannon's limit.

5.9 Phase-shift detection

Spread spectrum systems are sometimes used for covert systems. This is known as electronic countermeasures and provides decreased vulnerability to detection. Electronic counter-countermeasure (ECCM) receivers are designed to detect these types of signals. One of the methods to detect a BPSK waveform, as mentioned before, is to use a squaring, doubler, or $\times 2$ (doubling the frequency) to eliminate the phase shift as follows:

$$[A\cos(\omega t + 0°,180°)]^2 = A^2/2\cos(2\omega t + 0°,360°) = A^2/2\cos(2\omega t)$$

Note that $2 \times 0° = 0°$ and $2 \times 180° = 360°$, which is equal to $0°$. This is the basis for squaring, to eliminate the phase shift modulation so that the resultant signal is a CW spectral line instead of a sinc function with a suppressed carrier. The CW frequency is then easily detected, since this squaring process despreads the wideband signal.

The end result is a spectral line at twice the carrier frequency. This basic principle for detecting a BPSK data stream allows the ECCM receiver to obtain the frequency of the signal being sent by dividing the output frequency by two for BPSK. If BPSK is used, the null-to-null bandwidth, which is equal to twice the chip rate, allows the ECCM receiver to calculate the chip rate of the BPSK signal. The sidelobes are monitored to ensure that they are half of the width of the main lobe, which equals the chip rate. If they are not, then this provides additional information to the ECCM receiver that the signal is not BPSK and might be another form of modulation.

QPSK and offset QPSK (OQPSK) are detected by what is known as a times-4 ($\times 4$) detector, which quadruples the input signal to eliminate the phase ambiguities. OQPSK has all the same absolute phase states but is not allowed to switch more than $90°$, which eliminates the $180°$ phase shifts. However, the criteria depend on only the absolute phase states and how they are eliminated. The following shows the results of squaring the signal twice. The first squaring function produces:

$$\begin{aligned}[A\cos(\omega t+0°,90°,180°,-90°)]^2 &= A^2/2\cos(2\omega t+0°,180°,360°,-180°)\\ &= A^2/2\cos(2\omega t+0°,180°)\ \text{plus DC offset filtered out}\end{aligned}$$

Squaring again produces:

$$\begin{aligned}[A^2/2\cos(2\omega t + 0°, 180°)]^2 &= A^4/8\cos(4\omega t + 0°, 360°)\ \text{plus a DC term filtered out}\\ &= A^4/8\cos(4\omega t)\end{aligned}$$

Note: Since the possible phase states are 0°, 180°, 90°, and −90°, squaring them would only give $2 \times 0° = 0°$, $2 \times 180° = 360° = 0°$, $2 \times 90° = 180°$, and $2 \times -90° = -180° = 180°$. Therefore, the problem has been reduced to simple phase shifts of a BPSK level. One more squaring will result in the same as described in the previous BPSK example, which eliminates the phase shift. Thus, quadrupling or squaring the signal twice eliminates the phase shift for a quadrature phase-shifted signal.

The resultant signal gives a spectral line at four times the carrier frequency. This is the basic principle behind the ECCM receiver in detecting a QPSK or OQPSK data stream and allows the ECCM receiver to know the frequency of the signal being sent by dividing the output frequency by four for both QPSK and OQPSK waveforms.

For minimum shift keying (MSK), there is a sinusoidal modulating frequency proportional to the chip rate in addition to the carrier frequency along with the phase transitions. One way to generate classic MSK is to use two BPSK in an OQPSK-type system and sinusoidally modulate the quadrature channels at a frequency proportional to the bit rate before summation. In other words, the phase transitions are smoothed out by sinusoidal weighting. By positioning this into a times-4 detector as before, the resultant is:

$$\begin{aligned}&[A\cos(\omega t + 0°, 90°, 180°, -90°)(B\cos \pi t/2T)]^4\\&\quad = [AB/2(\cos(\omega \pm \pi/2T)t + 0°, 90°, 180°, -90°)]^4\end{aligned}$$

The phase ambiguities will be eliminated because the signal is quadrupled so that the phase is multiplied by 4. This results in 0°, so these terms are not carried out.

Squaring the equation first produces the sums and differences of the frequencies. Assuming that the carrier frequency is much larger than the modulating frequency, only the sum terms are considered and the modulating terms are filtered out. For simplicity, the amplitude coefficients are left out. Therefore, squaring the above with the conditions stated results in:

$$\cos(2\omega \pm \pi/T)t + \cos(2\omega)t$$

Squaring this equation with the same assumptions result in quadrupling MSK:

$$\cos 4\omega t + \cos(4\omega \pm 2\pi/T)t + \cos(4\omega \pm \pi/T)t$$

There will be spectral lines at four times the carrier, four times the carrier ± four times the modulating frequency, and four times the carrier ± two times the modulating frequency. Therefore, by quadrupling the MSK signal, the carrier can be detected. Careful detection can also produce the chip rate features since the modulating frequency is proportional to the chip rate.

5.10 Summary

The demodulation process is an important aspect in the design of the transceiver. Proper design of the demodulation section can enhance the sensitivity and

performance of data detection. Carrier recovery loops, such as the squaring loop, Costas loop, modified Costas loop, and decision directed Costas loop provide a means for the demodulator to strip off the carrier. Two types of demodulation can be used to despread and recover the data. The matched filter approach simply delays and correlates each delay segment of the signal to produce the demodulated output. This process includes the use of PPM to encode and decode the actual data. Another demodulation process uses a coherent sliding correlator to despread the data. This process requires alignment of the codes in the receiver, which is generally accomplished by a short acquisition code. Tracking loops, such as the early-late gate, align the code for the despreading process. A symbol synchronizer is required to sample the data at the proper time in the eye pattern to minimize the effects of ISI or use digital processing by taking multiple samples to determine the bit decision. Use of scramblers/descramblers to prevent long runs of ones or zeros, Shannon's limit for maximum throughput, and receivers designed for intercepting transmissions of other transmitters use various means of detection depending on the type of phase modulation utilized.

5.11 Problems

1. Which two ways are used to eliminate the carrier from the PSK digital waveform?
2. Given a BPSK signal, show how a squaring loop eliminates the phase ambiguity.
3. Using the squaring loop approach for BPSK detection, what would be needed in (a) an 8-PSK detector and (b) a 16-PSK detector?
4. What method is used to eliminate the carrier from a CP-PSK quadrature waveform?
5. What are the two basic ways to demodulate spread spectrum?
6. What is a matched filter correlator and how does it eliminate the spread spectrum code?
7. What is the null-to-null bandwidth of a pulse modulation which a pulse width of 1 μs?
8. What is PPM? What are the advantages and disadvantages between coherent and differential systems?
9. What is a sliding correlator and what is it used for?
10. What is a method used to maintain lock using a sliding correlator?
11. How does the bandwidth change in (a) the pulsed matched filter correlator and (b) the sliding correlator?
12. What is an eye pattern?
13. Where on the eye pattern is the best place for sampling the signal for best performance? Where is the worst place?
14. What is constant group delay? What is the result if the group delay is not constant?
15. What is ISI and how does it affect the desired waveform?

16. What is a scrambler and why is it used?
17. What are the two parameters that establish Shannon's limit?
18. MSK can be considered as a frequency-shifted signal, what would be another way of detection of MSK?

Further reading

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Holmes, Jack K. *Coherent Spread Spectrum Systems*. Hoboken, NJ: Wiley & Sons, 1982, pp. 251–267.

Keate, Christopher R. "The Development and Simulation of a Single and Variable Data-Rate Data-Aided Tracking Receiver." Ph.D. dissertation, 1986.

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Chapter 6
Basic probability and pulse theory

To achieve a better understanding of digital communications, some basic principles of probability theory need to be discussed. This chapter provides an overview of theory necessary for the design of digital communications and spread spectrum systems. Probability is used to calculate the link budget in regard to the error and required signal-to-noise ratio (SNR) and to determine whether a transceiver is going to work and at what distances. This is specified for digital communications systems as the required E_b/N_o.

6.1 Basic probability concepts

The central question concerning probability is whether something is going to occur, whether it is an error in the system, or the probability that multipath will prevent the signal from arriving at the receiver. The probability that an event occurs is described by the probability density function (PDF) and is defined as

$$f_x(x)$$

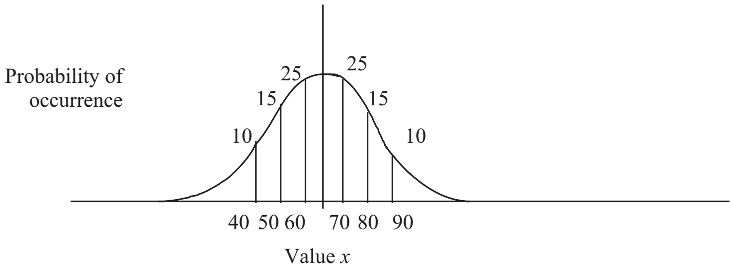
Probability can be expressed as a percentage, for example, a 10% chance that a value is present or that an event has occurred, or the probability of occurrence (Figure 6.1). The integral of the PDF is equal to $F_x(x)$ and is referred to as the cumulative distribution function (CDF). This integral of the PDF over $\pm\infty$ equals 1, which represents 100%.

For example, if there is a 10% chance of getting it right, by default, then there is a 90% chance of getting it wrong. The CDF is shown in Figure 6.2.

Often the PDF is mislabeled as the CDF. For example, if something has a Gaussian distribution, the curve that comes to mind is the PDF, as shown in Figure 6.1. This is not the CDF.

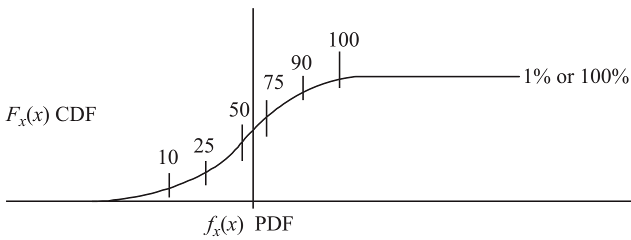
Another term used in probability theory is called the expected value $E[x]$. The expected value is the best guess as to what the value will be. To obtain the expected value, the signal values are multiplied by the percentage or the PDF and then summed or integrated as shown:

$$E[x] = \int x f_x(x)dx$$



Probability density function for Gaussian distribution

Figure 6.1 Probability density function for a Gaussian process



Cumulative distribution function for Gaussian distribution

Figure 6.2 Cumulative distribution function for a Gaussian process

The following is a discrete example where the density function is symmetrical. To plug in some numbers:

x	$f_x(x)$	$xf_x(x)$
1	0.1(10%)	0.1
2	0.2(20%)	0.4
3	0.4(40%)	1.2
4	0.2(20%)	0.8
5	0.1(10%)	0.5

$$\sum xf_x(x) = E[X] = 3.0$$

The mean (m_x) is equal to the expected value. Therefore;

$$m_x = E[x] = \int xf_x(x)dx = 3.0$$

6.2 The Gaussian process

The Gaussian process or distribution is probably the most common distribution in analyzing digital communications. There are many other types of distributions,

such as the uniform distribution, which is used for equal probability situations (e.g., the phase of a multipath signal), and the Rayleigh distribution, which is used to characterize the amplitude variations of the multipath. The Gaussian distribution, however, is used most frequently. It is called the normal distribution or bell-shaped distribution because of its common use and because it produces a PDF bell-shaped curve. Most often, noise is characterized using a Gaussian distribution for transceiver performance and SNR evaluations. The PDF for a Gaussian process is defined as follows:

$$f_x(x) = \frac{1}{\sigma\sqrt{2\pi}}\, e^{\frac{-(x-m_x)^2}{2\sigma^2}}$$

where σ is the standard deviation and m_x is the mean.

This establishes a curve called the distribution function. This is really the density function, a slight misnomer. Note that it is not the CDF as mentioned earlier. The cumulative distribution is defined as:

$$F_x(x) = \frac{1}{2}\left(1 + \mathrm{erf}\,\frac{x}{\sigma\sqrt{2}}\right)$$

where erf is the error function and σ is the standard deviation. The mean is assumed to be zero.

The erf value can be calculated, or a lookup table can be used with an approximation, if x is large. The CDF is used to calculate the percentage that the error is within a given range, such as 1σ or 2σ.

For example, what is the probability that x is between $\pm 2\sigma$ or a 2σ variation or 2 times the standard deviation. The cumulative distribution is used, and substituting -2σ for x in the previous equation yields:

$$F_x(x) = \frac{1}{2}\left(1 + \mathrm{erf}\,\frac{-2}{\sqrt{2}}\right) = \frac{1}{2}\left(1 + \mathrm{erf}(-\sqrt{2})\right)$$

Note that $\mathrm{erf}(-x) = -\mathrm{erf}(x)$. Therefore, $\mathrm{erf}(-2\quad 1/2) = -\mathrm{erf}(1.414) = -0.954$ (Table 6.1).

Finishing the calculations:

$$F_x(x) = 1/2(1 + (-0.954)) = 0.023$$

This represents the probability of the function at -2σ or less, as shown in Figure 6.3. Therefore, assuming Gaussian distribution with zero mean, the probability of the function being a $+2\sigma$ or greater is the same.

The probability of the signal being outside $\pm 2\sigma$ is 0.0456. Thus, the probability of x being inside these limits is $1 - 0.0456 = 0.954$, or 95.4%. This is the same result as simply taking the previously given erf function, $\mathrm{erf}\,(2)^{1/2} = 0.954$, or 95.4%. For a -1σ error, the error function of -0.707 is -0.682. The distribution function is 0.159, as shown in Figure 6.3. So for $\pm 1\sigma$, the probability of being outside these limits is 0.318. The probability of being within these limits is

Table 6.1 Error function erf(x)

	Hundredths digit of *x*									
x	**0**	**1**	**2**	**3**	**4**	**5**	**6**	**7**	**8**	**9**
0.0	0.00000	0.01128	0.02256	0.03384	0.04511	0.05637	0.06762	0.07886	0.09008	0.10128
0.1	0.11246	0.12362	0.13476	0.14587	0.15695	0.16800	0.17901	0.18999	0.20094	0.21184
0.2	0.22270	0.23352	0.24430	0.25502	0.26570	0.27633	0.28690	0.29742	0.30788	0.31828
0.3	0.32863	0.33891	0.34913	0.35928	0.36936	0.37938	0.38933	0.39921	0.40901	0.41874
0.4	0.42839	0.43797	0.44747	0.45689	0.46623	0.47548	0.48466	0.49375	0.50275	0.51167
0.5	0.52050	0.52924	0.53790	0.54646	0.55494	0.56332	0.57162	0.57982	0.58792	0.59594
0.6	0.60386	0.61168	0.61941	0.62705	0.63459	0.64203	0.64938	0.65663	0.66378	0.67084
0.7	0.67780	0.68467	0.69143	0.69810	0.70468	0.71116	0.71754	0.72382	0.73001	0.73610
0.8	0.74210	0.74800	0.75381	0.75952	0.76514	0.77067	0.77610	0.78144	0.78669	0.79184
0.9	0.79691	0.80188	0.80677	0.81156	0.81627	0.82089	0.82542	0.82987	0.83423	0.83851
1.0	0.84270	0.84681	0.85084	0.85478	0.85865	0.86244	0.86614	0.86977	0.87333	0.87680
1.1	0.88021	0.88353	0.88679	0.88997	0.89308	0.89612	0.89910	0.90200	0.90484	0.90761
1.2	0.91031	0.91296	0.91553	0.91805	0.92051	0.92290	0.92524	0.92751	0.92973	0.93190
1.3	0.93401	0.93606	0.93807	0.94002	0.94191	0.94376	0.94556	0.94731	0.94902	0.95067
1.4	0.95229	0.95385	0.95538	0.95686	0.95830	0.95970	0.96105	0.96237	0.96365	0.96490
1.5	0.96611	0.96728	0.96841	0.96952	0.97059	0.97162	0.97263	0.97360	0.97455	0.97546
1.6	0.97635	0.97721	0.97804	0.97884	0.97962	0.98038	0.98110	0.98181	0.98249	0.98315
1.7	0.98379	0.98441	0.98500	0.98558	0.98613	0.98667	0.98719	0.98769	0.98817	0.98864
1.8	0.98909	0.98952	0.98994	0.99035	0.99074	0.99111	0.99147	0.99182	0.99216	0.99248
1.9	0.99279	0.99309	0.99338	0.99366	0.99392	0.99418	0.99443	0.99466	0.99489	0.99511
2.0	0.99532	0.99552	0.99572	0.99591	0.99609	0.99626	0.99642	0.99658	0.99673	0.99688
2.1	0.99702	0.99715	0.99728	0.99741	0.99753	0.99764	0.99775	0.99785	0.99795	0.99805
2.2	0.99814	0.99822	0.99831	0.99839	0.99846	0.99854	0.99861	0.99867	0.99874	0.99880
2.3	0.99886	0.99891	0.99897	0.99902	0.99906	0.99911	0.99915	0.99920	0.99924	0.99928
2.4	0.99931	0.99935	0.99938	0.99941	0.99944	0.99947	0.99950	0.99952	0.99955	0.99957
2.5	0.99959	0.99961	0.99963	0.99965	0.99967	0.99969	0.99971	0.99972	0.99974	0.99975
2.6	0.99976	0.99978	0.99979	0.99980	0.99981	0.99982	0.99983	0.99984	0.99985	0.99986
2.7	0.99987	0.99987	0.99988	0.99989	0.99989	0.99990	0.99991	0.99991	0.99992	0.99992
2.8	0.99992	0.99993	0.99993	0.99994	0.99994	0.99994	0.99995	0.99995	0.99995	0.99996
2.9	0.99996	0.99996	0.99996	0.99997	0.99997	0.99997	0.99997	0.99997	0.99997	0.99998
3.0	0.99998	0.99998	0.99998	0.99998	0.99998	0.99998	0.99998	0.99999	0.99999	0.99999
3.1	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999
3.2	0.99999	0.99999	0.99999	1.00000	1.00000	1.00000	1.00000	1.00000	1.00000	1.00000

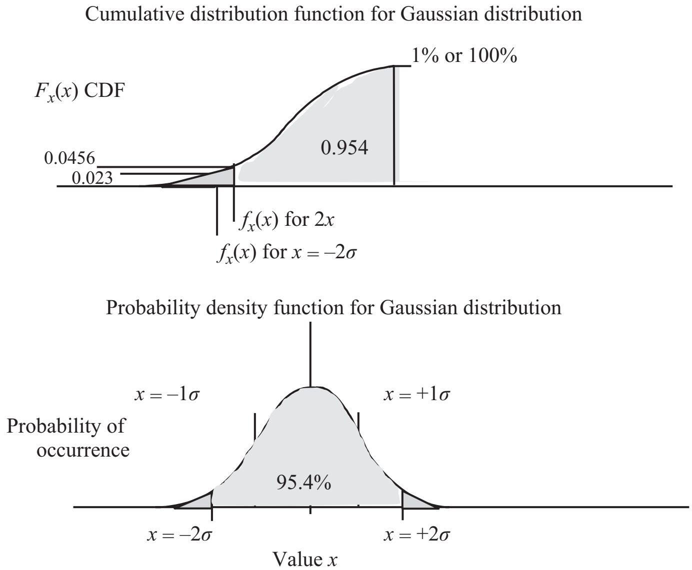


Figure 6.3 Probability within a given range using Gaussian distribution

$1 - 0.318 = 0.682$, or 68.2%. Given the range of x, the probability that the solution falls within the range is easily calculated.

Note that if the number of samples is not infinite, then the Gaussian limit may not be accurate and will degrade according to the number of samples that are taken.

6.3 Quantization and sampling errors

Since all digital systems are discrete and not continuous, the digitizing process creates errors between the digital samples. This error is created because a digitizing circuit samples the signal at regular intervals, resulting in a sequence of numerical values that simulates a continuous signal. If a signal is changing continuously with time, then, an error is caused by the estimation of the signal between the sample points. If the sample rate is increased, this estimation error is reduced.

There are basically two types of errors going from the analog signal to the digital signal: quantization error and sampling error. Quantization error occurs in the resolution of the analog signal due to discrete steps. This is dependent on the number of bits in the analog-to-digital converter: the more bits, the less quantization error. The value of the analog signal is now discrete and is either at one bit position or another. An example of quantization error is shown in Figure 6.4.

Sampling error comes about because the samples are noncontinuous samples. Basically, the sampling function samples at a point in time and extrapolates an estimate of the continuous analog signal from the points. This is dependent on the sample rate.

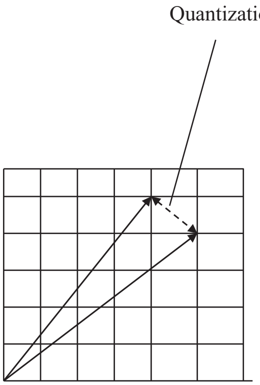


Figure 6.4 Quantization error in the A/D converter

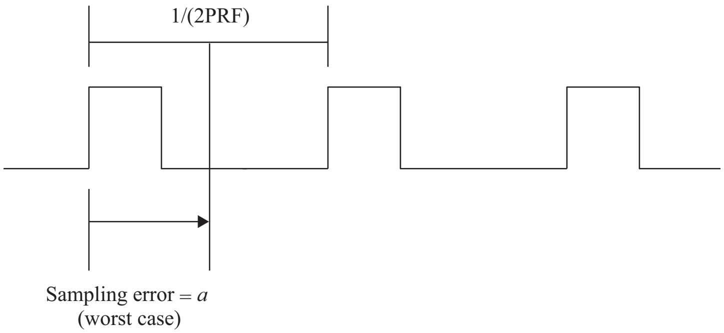


Figure 6.5 Sampling error showing worst case conditions

For example, radar range is based on the time of arrival of the pulse that is sent and returned. This is the basic principle that ranging and tracking radars work on when calculating the range of an aircraft from the ground station. One source of range error is called sampling error. If the arrival pulse is sampled, the sampling error is based on the distance between the sample pulses; halfway on either side is the worst case sampling error, as shown in Figure 6.5.

Since the probability that this occurrence is equal across the worst case points, it is referred to as being uniform or having a uniform distribution. The variance for a uniform distribution is:

$$\sigma^2 = a^2/3$$

where distribution varies between $+a$ and $-a$.

The standard deviation is equal to the square root of the variance:

$$\sigma = \sqrt{\frac{a^2}{3}}$$

For example, if the clock rate is 50 MHz, the time between pulses is 20 ns, so a, or the peak deviation, is equal to 10 ns. Therefore, the standard deviation is equal to

$$\sigma = \sqrt{\frac{10^2}{3}} = 5.77 \text{ ns}$$

If a number of samples are taken and averaged out, or integrated, then, the standard deviation is reduced by:

$$\sigma_{\text{ave}} = \frac{\sigma}{\sqrt{n_{\text{samples}}}}$$

If the number of samples that are to be averaged is nine, then, the standard deviation is:

$$\sigma_{\text{ave}} = \frac{5.77 \text{ ns}}{\sqrt{9}} = 1.92 \text{ ns}$$

To determine the overall error, the distribution from each source of error needs to be combined using a root sum square solution. This is done by squaring each of the standard deviations, summing, and then taking the square root of the result for the final error. For example, if one error has a $\sigma = 1.92$, and another independent error has a $\sigma = 1.45$, then, the overall error is:

$$\text{Total error } \sigma = \sqrt{1.92^2 + 1.45^2} = 2.41 \text{ ns}$$

This combination of the uniform distributions results in a normal or Gaussian distribution, overall in accordance with the central limit theorem. This is one of the reasons that the Gaussian or normal distribution is most commonly used. Therefore, even though each error is uniform and is analyzed using the uniform distribution, the resultant error is Gaussian and follows the Gaussian distribution for analysis. However, this assumes independent sources of error. In other words, one source of error cannot be related or dependent on another source of error.

6.4 Probability of error

The performance of the demodulation either is measured as the bit error rate (BER) or is predicted using the probability of error (P_e) as shown for binary phase-shift keying (BPSK):

$$\text{BER} = \text{Err}/\text{TNB}$$

$$P_e = \frac{1}{2}\text{erfc}\left(\frac{E_b}{N_o}\right)^{\frac{1}{2}}$$

where Err = number of bit errors, TNB = number of total bits, P_e = probability of error, E_b = energy in a bit, N_o = noise power spectral density (noise power in a 1-Hz bandwidth), erfc = 1 − erf = complimentary error function.

The BER is calculated by adding up the number of bits that were in error and dividing by the total number of bits for that particular measurement. BER counters continuously calculate this ratio. The BER could be called the bit error ratio since it is more related to a ratio than it is to time. An instrument used to measure or calculate BER is a BER tester.

Normally, the probability of error (P_e) is calculated from an estimated E_b/N_o for a given type of modulation. The probability of getting a zero when actually a one was sent for a BPSK modulated signal is:

$$P_{e1} = 1/2\text{erfc}((E_b/N_o)^{1/2})$$

The probability of getting a one when a zero was sent is:

$$P_{e0} = 1/2\text{erfc}\left((E_b/N_o)^{1/2}\right)$$

The average of these two probabilities is:

$$P_e = 1/2\text{erfc}\left((E_b/N_o)^{1/2}\right)$$

Note that the average probability is the same as each of the individual probabilities. This is because the probabilities are the same and are independent (i.e., either a one was sent and interpreted as a zero or a zero was sent and interpreted as a one). The probabilities happen at different times. The P_e for other systems are as follows:

$$\begin{aligned}
&\text{Coherent FSK}: P_e = 1/2\text{erfc}\left((E_b/2N_o)^{1/2}\right) \\
&\text{Noncoherent FSK}: P_e = 1/2\exp(-E_b/2N_o) \\
&\text{DPSK (noncoherent)}: P_e = 1/2\exp(-E_b/N_o) \\
&\text{Coherent quadrature phase shift keying (QPSK) and coherent minimum} \\
&\qquad \text{shift keying}: P_e = 1/2\text{erfc}\left((E_b/N_o)^{1/2}\right) - 1/4\text{erfc}^2\left((E_b/N_o)^{1/2}\right) \\
&\textit{Note}: \text{The second term can be eliminated for } E_b/N_o \gg 1
\end{aligned}$$

A spreadsheet with these plotted curves is shown in Figure 6.6. Note that the coherent systems provide a better P_e for a given E_b/N_o, or less E_b/N_o is required for the same P_e. Also note that coherent BPSK and coherent QPSK are close to the same at higher signal levels with a slight divergence with very small E_b/N_o ratios.

6.5 Error detection and correction

Error detection and correction are used to improve the integrity and continuity of service for a system. The integrity is concerned with how reliable the received information is. In other words, if information is received, how accurate are the data, and is it data or noise? Error detection can be used to inform the system that the data are false.

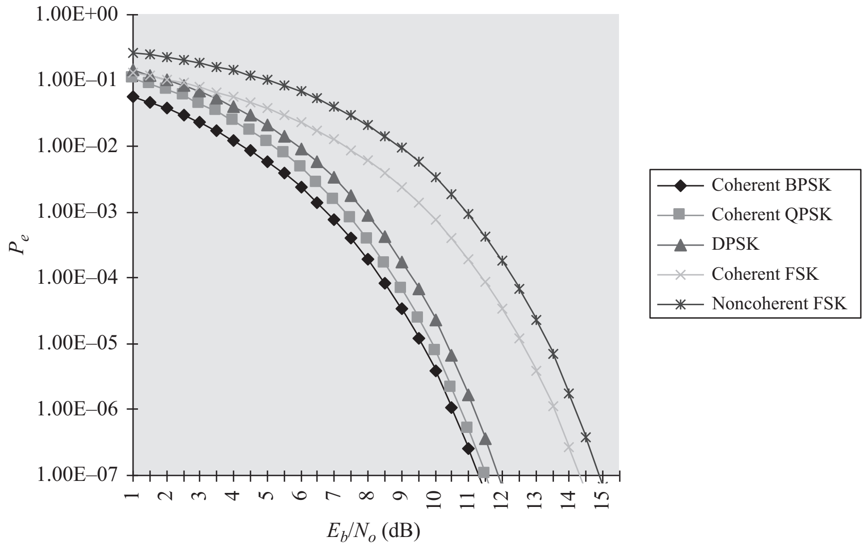


Figure 6.6 Probability of error curves for a given E_b/N_o

Continuity of service is concerned with how often the data are missed, which disrupts the continuity of the data stream. Error correction is used to fill in the missing or corrupted data to enhance the continuity of service.

Many techniques are used to perform both error detection and error correction. When designing a transceiver, the requirements need to be known as well as the amount of overhead the system can tolerate. Note that error detection and error correction require extra bits to be sent that are not data bits. So the amount of error detection and correction can significantly reduce the data capacity of a system, which is a trade-off in the system design. The extra bits are often referred to as overhead.

6.5.1 Error detection

There are several different methods of detecting errors in a system and informing the receiver that the information received from the transmitter is in error. If an error is detected, then the receiver can either discard the information or send a request to the transmitter to resend the information.

Three main detection schemes are commonly used for error detection in digital communication systems today: parity, checksum, and cyclic redundancy check (CRC).

6.5.2 Error detection using parity

Parity is used to determine if a message contains errors. The parity bit is attached to the end of a digital message and is used to make the number of "1"s that are in the message either odd or even. For example, if the sent message is 11011 and odd

parity is being used, then, the parity bit would be a "1" to make the number of "1"s in the message an odd number. The transmitter would then send 110111. If even parity is used, then, the parity bit would be a "0" to ensure that the number of "1"s is even, so the sent digital signal would be 110110. If the digital message is 11010, then, the sent digital signal would be 110100 for odd parity and 110101 for even parity.

To generate the parity bit, a simple exclusive-OR function of all the bits in the message including an extra bit produces the correct parity bit to be sent. The extra bit equals a "1" for odd parity and a "0" for even parity. For example, if the message is 11011, and the system is using odd parity (extra bit equals 1), then, the parity bit would be:

$$1 \text{ XOR } 1 \text{ XOR } 0 \text{ XOR } 1 \text{ XOR } 1 \text{ XOR } 1 \text{ (extra bit)} = 1$$

XOR is an exclusive-OR function in the digital domain. Consequently, the digital signal that is sent for this message would be 110111.

The exclusive-OR function is best described in the form of a truth table. When two bits are XORed, the truth table yields the resultant bit value. An XOR truth table is as follows:

Bit 1		Bit 2	Resultant bit value
0	XOR	0	0
0	XOR	1	1
1	XOR	0	1
1	XOR	1	0

The receiver detects an error if the wrong parity is detected. If odd parity is used, all of the received bits are exclusive-ORed, and if the result is a "1," then, there were no errors detected. If the result is a "0," then, this constitutes an error, known as a parity error. If even parity is used, then a "0" constitutes no errors detected and a "1" means that there is a parity error. For example, using odd parity, the receiver detects 110011, and by exclusive-ORing all the received bits, the output is a "0," which means that the system has a parity error.

One of the major drawbacks to this type of error detection is multiple errors. If more than one error occurs, then, this method of error detection does not perform very well. If two errors occur, it would look as if no errors occurred. This is true for all even amounts of error.

6.5.3 Error detection using checksum

Another method of detecting errors in a system is referred to as checksum. As the name connotes, this process sums up the digital data characters as they are being sent. When it has a total, the least significant byte of the sum total is appended to the data stream that is being sent.

The receiver performs the same operation by summing the input digital data characters and then comparing the least significant byte with the received least significant byte. If they are the same, then no error is detected. If they are different,

then an error is detected. This is called a checksum error. For example, if the message contained the following characters:

$$110, 011, 111, 011, 110, 001, 111, 110$$

Then, the digital sum would be 100,**111**. The least significant byte is **111**, which is attached to the message, so the final digital signal sent is:

$$110, 011, 111, 011, 110, 001, 111, 110, \mathbf{111}$$

The receiver would perform the same summing operation with the incoming data without the checksum that was sent. If the least significant byte is 111, then there is no checksum error detected. If it is different, then there would be a checksum error. Note that the receiver knows the length of the transmitted message.

6.5.4 Error detection using CRC

One of the best methods for error detection is known as a CRC. This method uses division of the data polynomial $D(x)$ by the generator polynomial $G(x)$, with the remainder being generated by using an XOR function instead of subtraction. The final remainder is truncated to the size of the CRC and is attached to the message for error detection. The data polynomial is multiplied by the number of bits in the CRC code to provide enough placeholders for the CRC. In other words, the remainder has to be large enough to contain the size of the CRC.

The polynomial is simply a mathematical expression to calculate the value of a binary bit stream. For example, a bit stream of 10101 would have a polynomial of $X^4 + X^2 + X^0$. The numeric value is calculated by substituting $X = 2$: $2^4 + 2^2 + 2^0 = 21$. Therefore, if there is a "1" in the binary number, the value it represents is included in the polynomial. For example, the first one on the left in the aforementioned digital word has a value of 16 in binary. Since $2^4 = 16$, this is included in the polynomial as X^4.

The receiver performs the same operation as before using the complete message, including the attached error detection bits, and divides by the generator polynomial. If the remainder is zero, then, no errors were detected. For example, suppose, we have a data polynomial, which is the data to be sent, and a generator polynomial, which is selected for a particular application:

$$D(x) = X^6 + X^4 + X^3 + X^0 \ 1011001$$
$$G(x) = X^4 + X^3 + X^2 + X^0 \ 11101$$

To provide the required placeholders for the CRC, $D(x)$ is multiplied by X^4, which extends the data word by four places:

$$X^4\left(X^6 + X^4 + X^3 + X^0\right) = X^{10} + X^8 + X^7 + X^4 \ 1011001\mathbf{0000}$$

Note this process is equivalent to adding the number of zeros as placeholders to the data polynomial, since the generator polynomial is X^4 (4 zeros). The division uses XOR instead of subtraction (Figure 6.7). Therefore, the CRC would be 0110, and

```
           1 1 1 1 1 1 0
1 1 1 0 1 |1 0 1 1 0 0 1 0 0 0 0
           1 1 1 0 1
           0 1 0 1 1 0
             1 1 1 0 1
             0 1 0 1 1 1
               1 1 1 0 1
               0 1 0 1 0 0
                 1 1 1 0 1
                 0 1 0 0 1 0
                   1 1 1 0 1
                   0 1 1 1 1 0
                     1 1 1 0 1
                     0 0 0 1 1 0
                         CRC
```

Figure 6.7 CRC generator using division with XOR instead of subtraction

the digital signal that would be transmitted is the data plus the CRC, which equals 10110010110.

The receiver would receive all the data including the CRC and perform the same function (Figure 6.8). Since the remainder is zero, the CRC does not detect any errors. If there is a remainder, then, there are errors in the received signal. Some of the standard CRCs used in the industry are listed in Table 6.2.

6.5.5 Error correction

Error correction detects and also fixes errors or corrects the sent data. One form of error correction is to use redundancy. Even though it is not actually correcting the error that is sent, this method involves sending the data more than once. Another form of error correction, which is really a true form of error correction, is forward error correction (FEC).

6.5.6 Error correction using redundancy

Error correction using redundancy means correcting the errors by having the transmitter resend the data. There are basically two concepts for this type of correction.

The first method requires no feedback on the sent message. The transmitter merely sends the data more than once to increase the probability that the message will be received. For example, if a burst signal occurs on the first time the message is sent, then by sending the message again the probability is increased that the receiver will receive the message.

The second method requires feedback from the receiver. The transmitter sends the message to the receiver, and if the receiver detects an error, it sends out a request to the transmitter to resend the message. The process is referred to as automatic repeat request.

```
            1 1 1 1 1 1 0
1 1 1 0 1 | 1 0 1 1 0 0 1 0 1 1 0
            1 1 1 0 1
            0 1 0 1 1 0
              1 1 1 0 1
              0 1 0 1 1 1
                1 1 1 0 1
                0 1 0 1 0 0
                  1 1 1 0 1
                  0 1 0 0 1 1
                    1 1 1 0 1
                    0 1 1 1 0 1
                      1 1 1 0 1
                      0 0 0 0 0 0
                          No errors
```

Figure 6.8 CRC error detection using division with XOR instead of subtraction

Table 6.2 A list of common CRC generator polynomials

- CRC-5 USB token packets $G(x) = x^5 + x^2 + x^0$
- CRC-12 telecom systems $G(x) = x^{12} + x^{11} + x^3 + x^2 + x^1 + x^0$
- CRC-ANSI $G(x) = x^{16} + x^{15} + x^2 + x^0$
- CRC-CCITT $G(x) = x^{16} + x^{12} + x^5 + x^0$
- IBM-SDLC $= x^{16} + x^{15} + x^{13} + x^7 + x^4 + x^2 + x^1 + x^0$
- IEC TC57 $= x^{16} + x^{14} + x^{11} + x^8 + x^6 + x^5 + x^4 + x^0$
- CRC-32 IEEE Standard 802.3 $G(x) = x^{32} + x^{26} + x^{23} + x^{22} + x^{16} + x^{12} + x^{11} + x^{10} + x^8 + x^7 + x^5 + x^4 + x^2 + x^1 + x^0$
- CRC-32C (castagnoli) $G(x) = x^{32} + x^{28} + x^{27} + x^{26} + x^{25} + x^{23} + x^{22} + x^{20} + x^{19} + x^{18} + x^{14} + x^{13} + x^{11} + x^{10} + x^9 + x^8 + x^6 + x^0$
- CRC-64-ISO ISO 3309 $G(x) = x^{64} + x^4 + x^3 + x^1 + x^0$
- CRC-64 ECMA-182 $G(x) = x^{64} + x^{62} + x^{57} + x^{55} + x^{54} + x^{53} + x^{52} + x^{47} + x^{46} + x^{45} + x^{40} + x^{39} + x^{38} + x^{37} + x^{35} + x^{33} + x^{32} + x^{31} + x^{29} + x^{27} + x^{24} + x^{23} + x^{22} + x^{21} + x^{19} + x^{17} + x^{13} + x^{12} + x^{10} + x^9 + x^7 + x^4 + x^1 + x^0$
- CRC-16 cell phones $G(x) = x^{16} + x^{15} + x^5 + 1$
- Others

Both of these methods take time, which slows down the data rate for a system, making these rather inefficient means of correction or error recovery. Also, if an error occurs in every message that is transmitted, then these methods are not a viable means of error correction.

6.5.7 Forward error correction

FEC not only detects errors but also makes an attempt to correct the error so that the transmitter does not have to send the message again. There are several types of

FEC, from very simple methods that can correct one error to very complex methods that can correct multiple errors. A process known as interleaving can improve error-correction methods against burst errors.

6.5.8 Types of FEC

Two types of FEC are commonly used in communications: block codes and convolutional codes.

Block codes. These receive k information bits into a block encoder, which is memoryless (does not depend on past events). The block encoder maps these information bits into an n-symbol output block with a rate $R = k/n$ (Figure 6.9). Block codes are specified as (n,k), where n is the number of bits in the output code word, and k is the number of information bits or data bits in the data word. For example, a block code of (7,4) would have 7 bits in the output code word and 4 bits in the data code word. Thus, each 4-bit information/data code word is mapped into a 7-bit symbol code word output. This means there are 2^k number of distinct 4-bit input messages, which also means that there are 2^k number of distinct 7-bit output code words. The output code is called a block code. The number of bits in the output code word is always greater than the number of bits in the data word: $n > k$.

For long block codes, 2^k with length n is complex and can take up a lot of memory. To reduce the amount of memory needed for these code words, linear block codes are used. They are easier to encode because they store only the generator matrix and do a linear combination of the generator matrix and the data word; in other words, they perform a modulo-2 sum of two linear block code words, which produces another linear block code word (Table 6.3).

Many codes are systematic and contain the actual information bits plus some parity bits. For a (7,4) linear block code, the code word contains the 4-bit data word plus 3 bits of parity (Table 6.3).

Error detection for linear systematic block codes calculates the syndrome. The syndrome is equal to the received matrix r times the transpose of the parity check matrix H^{-T}:

$$s = r \times H^{-T}$$

If the syndrome is equal to zero, then, there are no errors. Table 6.4 shows the process used to generate the syndrome.

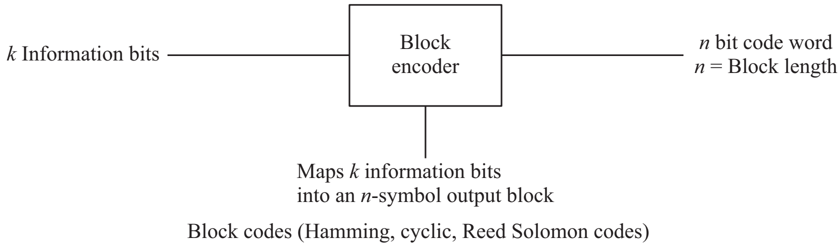


Figure 6.9 Block codes for FEC

Table 6.3 Linear systematic block code generator

Generator matrix includes identity matrix

Generator matrix	**Message**	**Modulo-2**
(**1000011**)	**1**	(1000011)
(**0100110**)	**0**	
(**0010111**)	**1**	(0010111)
(**0001101**)	**0**	
Systematic code word =		**(1010100)**
Identity matrix		
Messages	**Codewords**	**(7,4)**
(0000)	(0000000)	
(0001)	(0001101)	
(0010)	(0010111)	
(0011)	(0011010)	
(0100)	(0100110)	
(0101)	(0101011)	
(0110)	(0110001)	
(0111)	(0111100)	
(1000)	(1000011)	
(1001)	(1001110)	
(1010)	**(1010100)**	
(1011)	(1011001)	
(1100)	(1100101)	
(1101)	(1101000)	
(1110)	(1110010)	
(1111)	(1111111)	

XOR

4 bits of data **1010**
3 parity bits **100**

Table 6.4 Syndrome generator for linear systematic block codes

		$\underline{H^{-T}}$				
		0 1 1	$s = (s2, s1, s0)$			
$r = (r6, r5, r4, r3, r2, r1, r0)$ *		1 1 0		Modulo-2		
		1 1 1 =	$s2 =$	$r5 + r4 + r3 + r2$		
		1 0 1	$s1 =$	$r6 + r5 + r4 + r1$		XOR
	Identity	1 0 0	$s0 =$	$r6 + r4 + r3 + r0$		
	matrix	0 1 0	$s2, s1, s0 = 0{,}0{,}0$ no errors			
		0 0 1	**Example: $s2, s1, s0 = 0{,}0{,}1$**			

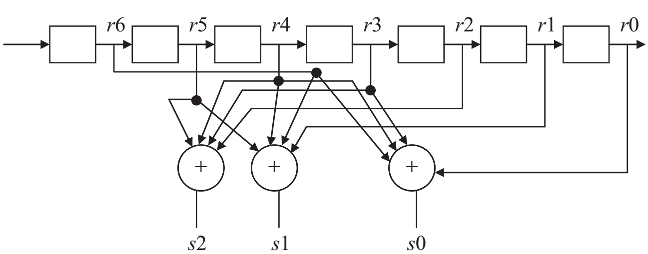


Table 6.5 Syndrome generator example, error detection for linear systematic block codes

Generator matrix	**Parity matrix *H***				
(1000011)	**(0111100)**				
(0100110)	**(1110010)**				
(0010111)	**(1011001)**				
(0001101)					
Identity matrix					
Code word received	H^{-T}	**Modulo-2**			
(1010100) *	0 1 1	0 1 1			
	1 1 0	0 0 0			
	1 1 1 =	1 1 1			
	1 0 1	0 0 0			
	1 0 0	1 0 0			
	0 1 0	0 0 0			
	0 0 1	0 0 0			
		0 0 0	Equal to zero, no errors		
Code word received	H^{-T}	**Modulo-2**			
(1010101) *	0 1 1	0 1 1			
	1 1 0	0 0 0			
	1 1 1 =	1 1 1			
	1 0 1	0 0 0			
	1 0 0	1 0 0			
	0 1 0	0 0 0			
	0 0 1	0 0 1			
		0 0 1	Not equal to zero, errors		

An example showing the generation of the syndrome with both the no-error case and the case where there is an error occurring in the last bit is shown in Table 6.5. The syndrome is equal to (0,0,0) for the no-error case and is equal to (0,0,1) for the error case. This is using the same (7,4) linear block code as before. Therefore, the error is detected with the syndrome not equal to (0,0,0).

A measure of the ability to distinguish different code words is called the minimum distance. The minimum distance between the code words is the number of bits that are different. To calculate the minimum distance between code words, the code words are modulo-2 added together, and the number of "1"s in the resultant is the minimum distance. For example,

Code word 1 = 1100101
Code word 2 = 1000011
Modulo-2 = 0100110 = three "1"s
Minimum distance = 3

This is also equal to the minimum weight for linear block codes, three "1"s in code word 2. To ensure that the minimum weight is accurate, all code possibilities need to be examined to determine the fewest number of "1"s in a code word.

The ability for an error-detection method to detect errors is based on the minimum distance calculated. The error-detection capability, or the number of errors that can be detected, is equal to:

$$\text{Number of errors detected} = d_{\min} - 1$$

where $d_{\min}$ is the minimum distance. This example shows that the number of errors that can be detected is equal to:

$$d_{\min} - 1 = 3 - 1 = 2$$

Even though the system is able to detect more errors equal to and greater than $d_{\min}$, it cannot detect all of the errors. Thus, undetected errors will pass through the system.

The ability of an error-correction method to correct errors is also based on the minimum distance calculated. The error-correction capability, or the number of errors that can be corrected, is equal to:

$$\text{Number of errors corrected} = (d_{\min} - 1)/2$$

This example shows that the number of errors that can be corrected is equal to:

$$\text{Number of errors corrected} = (d_{\min} - 1)/2 = (3 - 1)/2 = 1$$

An error-correction method can perform a combination of both error correction and detection simultaneously. For example, if $d_{\min}$ is equal to 10, the combined ability of this method could correct three errors and detect six errors or correct four errors and detect five. Error correction and detection codes often specify the minimum distance in their description:

$$(7{,}4{,}3) = (n{,}k{,}d_{\min})$$

An example of error correction is shown in Table 6.6. The error vector, which is the received vector that contains an error, is multiplied by the transpose of the parity check matrix H^{-T}. This generates three error equations equivalent to the syndrome calculated in Table 6.5, which is equal to (0,0,1). There are three equations and seven unknowns, so the possible solutions are equal to:

$$2^{(7-3)} = 2^4 = 16$$

These equations are solved to find the solution with the most zeros. The solutions with the most zeros are known as coset error vectors. The example shows that if all the e values are set to zero except for $e_0 = 1$, then, the coset error vector is equal to 0000001. To correct the code error, the received code is modulo-2 added to the coset error vector to correct the error:

$$\begin{aligned}&\text{Received code} = 1010101\\&\text{Coset error vector} = \underline{0000001}\\&\text{Correct code} = 1010100\end{aligned}$$

Table 6.6 Error vector and error correction example for linear systematic block codes

Error vector	H^{-T}				**Coset, error vectors**
$e_6,e_5,e_4,e_3,e_2,e_1,e_0$ *	0 1 1	0	e_6	e_6	1 0 0 0 0 0 0
	1 1 0	e_5	e_5	0	0 1 0 0 0 0 0
	1 1 1 =	e_4	e_4	e_4	0 0 1 0 0 0 0
	1 0 1	e_3	0	e_3	0 0 0 1 0 0 0
	1 0 0	e_2	0	0	0 0 0 0 1 0 0
	0 1 0	0	e_1	0	0 0 0 0 0 1 0
	0 0 1	0	0	e_0	**0 0 0 0 0 0 1**
		$e_5+e_4+e_3+e_2 = 0$			$\boldsymbol{e_6,e_5,e_4,e_3,e_2,e_1,e_0}$
			$e_6+e_5+e_4+e_1 = 0$		
				$e_6+e_4+e_3+e_0 = 1$	
	syndrome =	**0**	**0**	**1**	
3 equations, 7 unknowns (7−3=4), possible solutions = $2^4 = 16$					
Solve for a solution with the most zeros $e_6 = e_5 = e_4 = e_3 = e_2 = e_1 = 0$, $e_0 = 1$					
Therefore the coset, error vector = 0000001 satisfies the three above equations with most zeros					
The received code is 1010101 in error; the correct code sent was 1010100.					
$Vc = r+e =$	101010**1** + **0000001** = 101010**0** Corrected bit in code word				

Note: + = XOR

The probability of receiving an undetected error for block codes is calculated with the following equation:

$$P_u(E) = \sum_{i=1}^{n} A_i p^i (1-p)^{n-i}$$

where p is the probability of error, i is the number of "1"s in the code word, A_i is the weight distribution of the codes (number of code words with the number of "1"s specified by i).

An example is shown in Table 6.7 for a (7,4) block code. Note that there is one code word with no "1"s, one code word with seven "1"s, seven code words with three "1"s, seven code words with four "1"s, and no code words with one, two, five, and six "1"s. Therefore, this example shows that the probability of an undetected error is equal to:

$$P_u(E) = 7p^3(1-p)^4 + 7p^4(1-p)^3 + p^7$$

If the probability of error equals 10^{-2}, then, the probability of an undetected error is

$$\begin{aligned} P_u(E) &= 7(10^{-2})^3(1-(10^{-2}))^4 + 7(10^{-2})^4(1-(10^{-2}))^3 + (10^{-2})^7 \\ &= 6.8 \times 10^{-6} \end{aligned}$$

Table 6.7 Probability of undetected error performance of block codes

$P_u(E) = \sum_{i=1}^{n} A_i p^i (1-p)^{n-i}$

where p is the probability of error and A_i is the weight distribution of the code as shown below for a 7,4 code

Codewords			
(0000000)	n = #bits of code = 7		
(0001101)	i = # of "1"s		
(0010111)	A_i is the number of code words		
(0011010)	with the same number of "1"s		
(0100110)	A_0 = Not used, no ones		
(0101011)	$A_1 = 0$		
(0110001)	$A_2 = 0$		
(0111100)	$A_3 = 7$	7 codes have 3 ones	
(1000011)	$A_4 = 7$	7 codes have 4 ones	
(1001110)	$A_5 = 0$		
(1010100)	$A_6 = 0$		
(1011001)	$A_7 = 1$	1 code has 7 ones	
(1100101)			
(1101000)			
(1110010)			
(1111111)			

If the probability of error equals 10^{-6}, then, the probability of an undetected error is

$$\begin{aligned} P_u(E) &= 7(10^{-6})^3(1-(10^{-6}))^4 + 7(10^{-6})^4(1-(10^{-6}))^3 + (10^{-6})^7 \\ &= 7 \times 10^{-18} \end{aligned}$$

The *Hamming code* is another popular block code. The number of bits required for the Hamming code is:

$$2^n \geq m + n + 1$$

where n is the number of Hamming bits required, m is the number of data bits.

Hamming bits are randomly placed in the data message. For example, suppose, a digital data stream is 1011000100010, which is 13 bits long. The number of Hamming bits required is:

$$2^5 \geq 13 + 5 + 1$$

where $n = 5$. The Hamming bits slots are placed randomly in the data message as follows:

17					15			14				8				3	
H	1	0	1	1	H	0	0	H	0	1	0	H	0	0	1	H	0

The bit values of the Hamming bits are calculated by exclusive-ORing the bit positions that contain a "1," which are 3, 8, 14, 15, and 17:

00011	3
01000	8
01110	14
01111	15
10001	17
11011	Hamming Code

Consequently, the signal that would be sent out of the transmitter would be:

H					H			H				H				H	
1	1	0	1	1	**1**	0	0	**0**	0	1	0	**1**	0	0	1	**1**	0

If an error occurs at bit position 15, which means that the "1" value at the transmitted signal at bit 15 is detected as a "0" in the receiver, then, the following bit stream would be received:

110$\underline{0}$11000010100110

To detect and correct this error, at the receiver, all bits containing a "1" are exclusive-ORed:

00010	2
00011	3
00110	6
01000	8
01101	13
01110	14
10001	17
10010	18
01111	15

The result, $01111 = 15$, is equal to the bit position where the error occurred. To correct the error, bit 15 is changed from a "0" to a "1."

The Hamming block code corrects only one error. Other higher level codes, such as the Reed–Solomon code, have the ability to correct multiple errors. Some other types of codes used for error correction include:

- Hsiao code—uses a shortened Hamming code.
- Reed–Muller code—used for multiple error corrections.
- Golay code—used in space programs for data communications.

In addition, a very popular communications code is the cyclic code, which shifts one code word to produce another. This is a subset of linear block codes that uses linear sequential circuits such as shift registers with feedback to generate the codes. Since they are serially generated, they are able to be decoded serially. Reed–Solomon code is generated by cyclic codes and is used extensively for

communications. Bose–Chaudhuri–Hocquenghem code is another good example of a cyclic code. Even Hamming and Golay codes can be made to be cyclic codes.

Sometimes, shortened codes are used. This is where the data are zero padded (make some data symbols zero at the encoder); they are not transmitted, but they are reinserted at the decoder.

Error correction and detection methods improve the system's ability to detect and correct errors to ensure more reliable communications. However, the cost of doing this is either more bandwidth to send more bits at a higher rate or a decrease in data throughput to send more bits at the same rate. These additional bits required for error correction and detection are often referred to as overhead. Along with the overhead concerns with error detection and correction, another caution is that for high error rates (10^{-2}) or more error correction may actually degrade performance.

Convolutional codes: These codes are also commonly used because of their simplicity and high performance. They are usually generated using a pseudonoise (PN) code generator with different taps selected for the XOR functions, and a clock rate that selects output symbols generally at a higher rate than the input symbols. For example, suppose, the output rate is equal to twice the speed of the input information, this would produce a rate as follows:

$$R = \text{Input information bit rate/output symbol clock rate} = 1/2$$

Also, for convolutional codes, the size of the shift register is called the constraint length. Therefore, if the shift register is seven delays long, then, this example would be a convolutional code with rate 1/2 and constraint length 7 (Figure 6.10). Many different rates are possible, but some of the more common ones are 1/2, 3/4, and 7/8.

The more common methods of decoding use maximum likelihood sequence (MLS) estimation and the Viterbi algorithm, which is a modification of the standard MLS decoding scheme.

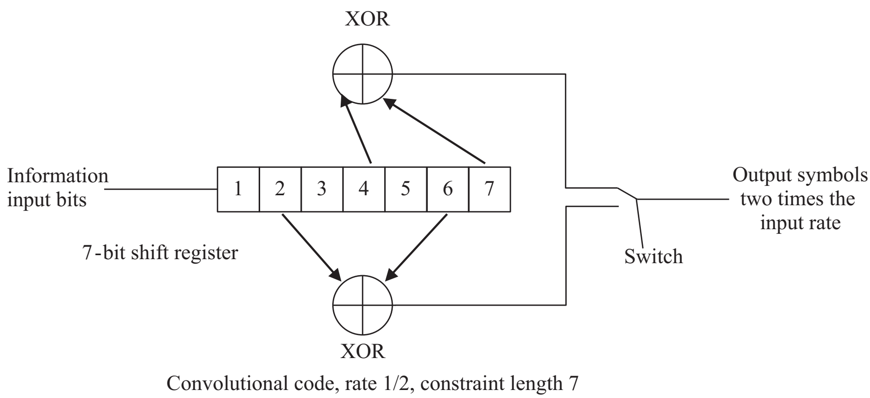


Figure 6.10 Convolutional codes for FEC

The MLS performs the following three basic operations:

1. It calculates the distances (the amount of difference, generally minimizes bit errors) between the symbol sequence and each possible code sequence.
2. It selects the code sequence that has the minimum distance calculated as being the correct code sequence.
3. It outputs this sequence as the MLS estimation of the received sequence.

Therefore, to retrieve the data information that was sent, the receiver performs the deinterleaving process and sorts out the data bits in the correct messages. Then, a decoder, such as a Viterbi decoder, is used to perform the necessary error correction to receive reliable data.

6.5.9 Interleaving

Large interfering signals may occur for a short period of time, or burst, causing several errors for that short period of time. Since most error-correcting schemes only correct a small percentage of errors in a data stream, they are unable to correct several errors at one time. To minimize this problem, the data are spread out and interleaved with other data signals so that when a burst occurs, instead of affecting one data message with several errors, only a small number of errors are present in each data message. Therefore, when the received data are deinterleaved, the error correction for each message contains a minimal number of errors to correct.

To illustrate interleaving, we start out with a digital signal using "a" as a placeholder for the bit value in a message. An example would be:

a a a a **<u>a a a</u>** a

Suppose, a burst jammer occurred that was wide enough to cause three of these bits to be in error (the underlined bits). If three digital messages needed to be sent, interleaving could be performed to allow only 1-bit error in message "a":

Message a Message b Message c
a a a **<u>a a a</u>** a a a b b b b b b b b c c c c c c c c
Interleaved Messages a, b, c
a b c **<u>a b c</u>** a b c a b c a b c a b c a b c a b c

The jamming signal still caused three errors at the same point in time; however, it caused only one error per message. The deinterleaving process is used to sort out the bits, and then, error correction is used to correct the bits in error for each message:

Received messages after deinterleaving:
a **<u>a</u>** a a a a a a b **<u>b</u>** b b b b b b c **<u>c</u>** c c c c c c

Therefore, instead of losing message "a," one error is corrected in each of the messages, and the correct data are obtained. This method of interleaving and deinterleaving the data is used extensively in systems that are susceptible to burst jammers or burst noise.

Two more examples of interleaving to reduce the FEC required 1/4 and 1/16 is shown in Figures 6.11 and 6.12.

6.5.10 Viterbi decoder

A Viterbi decoder is a real-time decoder for very high-speed short codes, with a constraint length of less than 10. It is a major simplification of the MLS estimator. The basic concept is that it makes real-time decisions at each node as the bits progress through the algorithm. Only two paths reach a node in the trellis diagram, and the Viterbi algorithm makes a decision about which of the paths is more likely to be correct (Figure 6.13). Only the closest path to the received sequence needs to be retained.

Multi-h is a technique to improve detection of a continuous phase modulation system by changing the modulation index. For example, if the modulation index changes between 1/4 and 1/2, often specified by $\{1, 2/4\}$, then, the modulation changes between $\pi/4$ and $\pi/2$. This means that the modulation changes from a $\pi/4$ shift to a $\pi/2$ shift, which improves the detection process by extending the ambiguity of the nodes of the trellis diagram. Therefore, the first merge point is twice as far on the multi-h trellis diagram (Figure 6.14). Extending the ambiguity points farther in the trellis diagram lessens the number of errors that can occur or the number of bit decisions.

The multi-h method improves spectral efficiency by providing more data throughput for a given bandwidth. The disadvantage is that the system becomes more complex. In addition, it has constraints that require:

$$q \geq M^k$$

where q is the common denominator of modulation index (4 in the example), M is the number of phase states (binary in the example), K is the number of modulation states (2 in the example). This example shows this meets the constraint for a good use of multi-h: $4 = 2^2$.

6.5.11 Turbo and low-density parity check codes

Two other FEC codes used in data link communications are *turbo codes* and *low-density parity check* (LDPC) *codes*. Turbo codes were invented in 1993 by Claude Berrou and Alain Glavieux. They use two or more parallel convolutional encoders/decoders. These codes claim that they are approximately 0.5 dB away from Shannon's limit. They use parallel-concatenated convolutional codes. The disadvantage of turbo codes is the decoding delay, since iterations take time, and this may affect real-time voice, hard disk storage, and optical transmissions.

Turbo codes use a recursive systematic code. It is recursive because it uses feedback of the outputs back into the input. It is systematic because one of the outputs is exactly the same as the input. These two properties provide better performance over nonrecursive, nonsystematic codes. The term turbo refers to turbo engines, which use exhaust to force air into the engine to boost combustion, similar

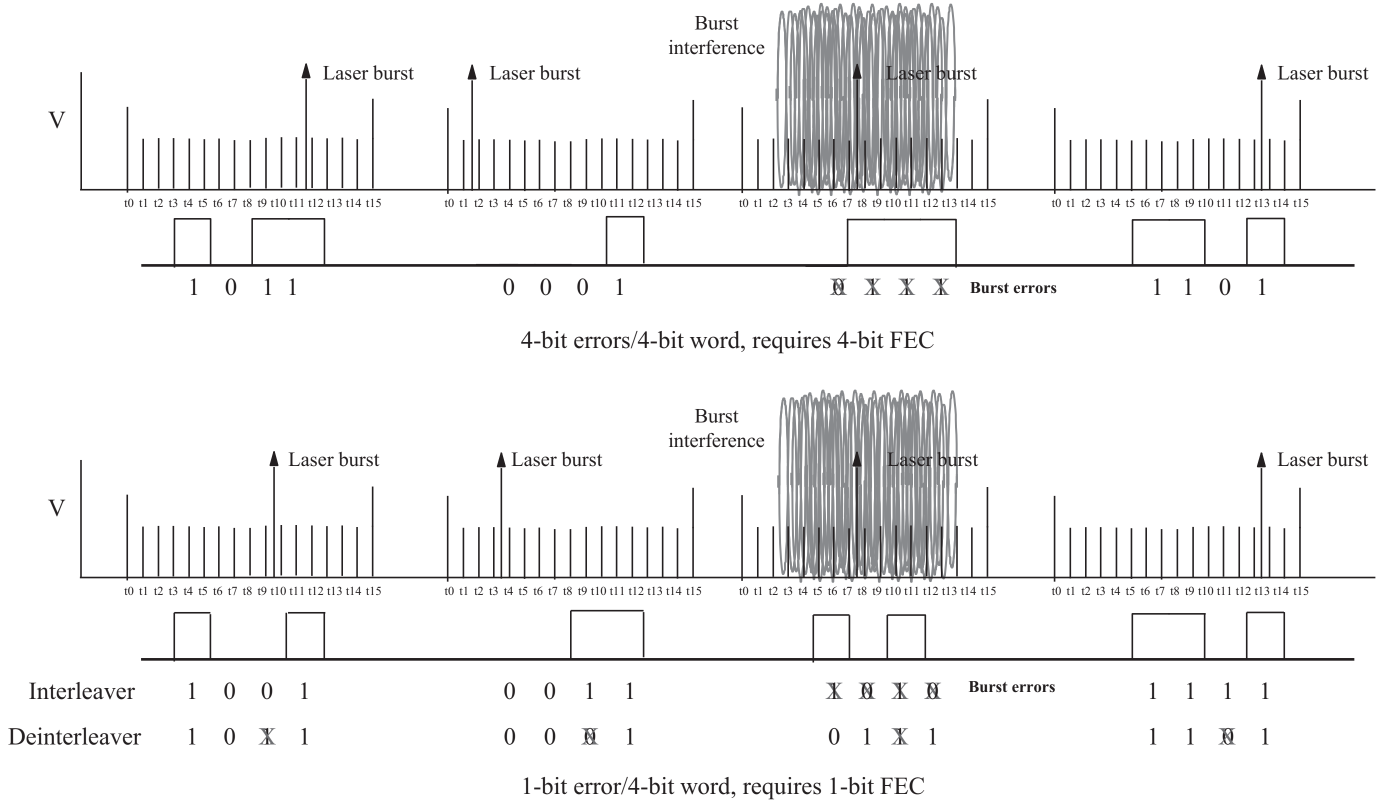


Figure 6.11 Interleaving reduces FEC required by 1/4 for burst errors

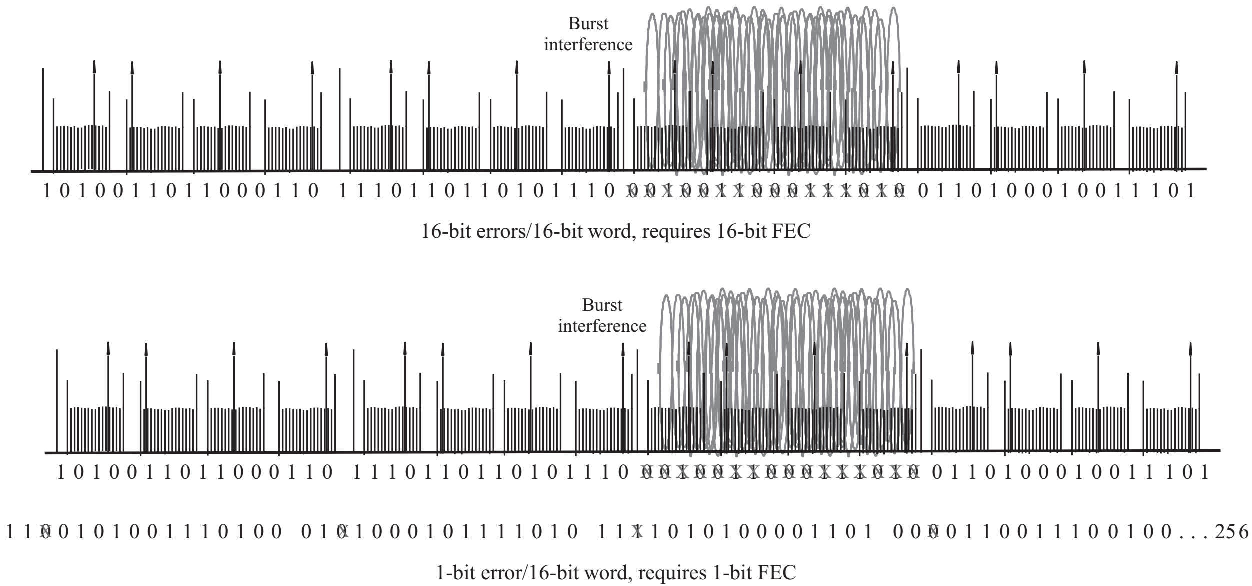


Figure 6.12 Interleaving reduces FEC required by 1/16 for burst errors

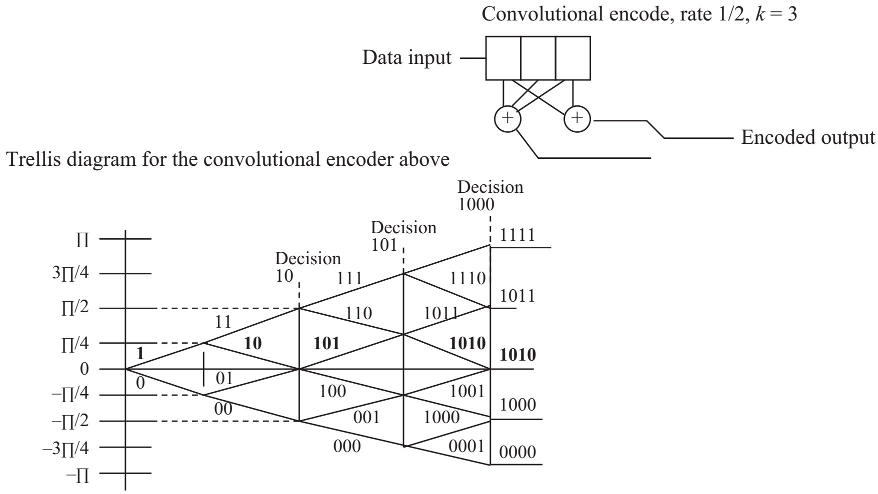


Figure 6.13 Viterbi decoder trellis diagram

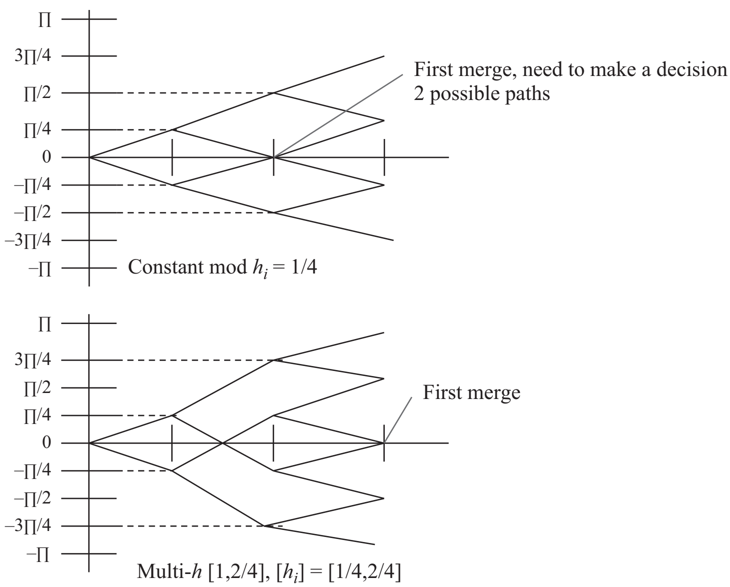


Figure 6.14 Trellis diagram for constant mod and multi-h

to the function of the code, where it provides feedback from the output of the decoder to the input of the decoder.

The transmitted code word structure is composed of the input bits, parity bits from the first encoder using the input bits, and parity bits from the second encoder

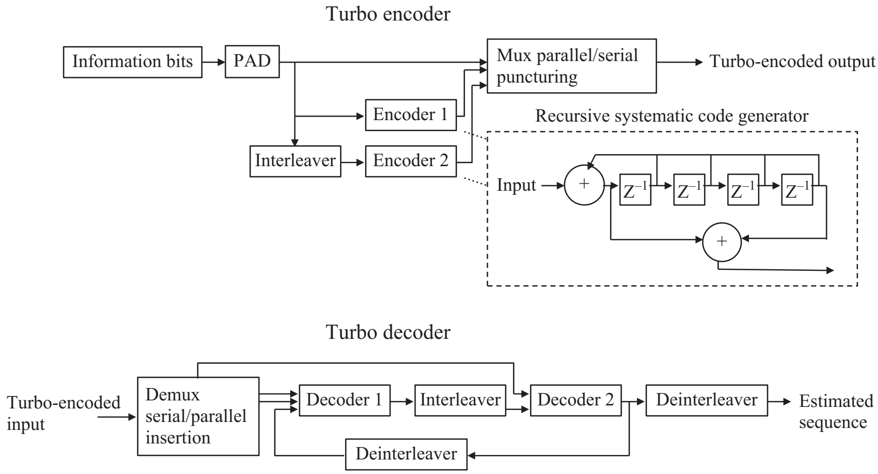


Figure 6.15 Basic Turbo code diagram

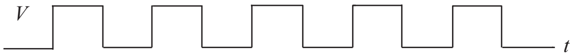
using interleaved input bits (Figure 6.15). The decoders use log-likelihood ratios as integers with reliability estimate numbers of -127 (0) to $+127$ (1). It uses an iterative process using both decoders in parallel.

A technique known as puncturing is also used. This process deletes some bits from the code word according to a puncturing matrix to increase the rate of the code. For example, puncturing can change the rate from 1/2 to 2/3, while still using the same decoder for these different rates. However, systematic bits (input bits) in general are not punctured. Puncturing provides a trade-off between code rate and performance.

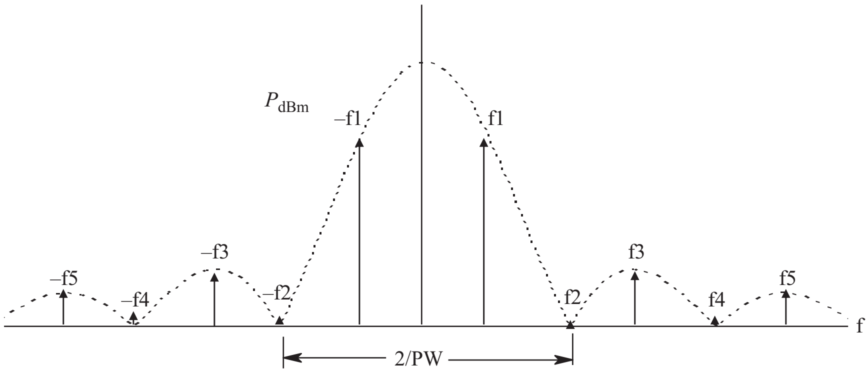
LDPC codes outperform turbo codes, with claims that the performance is approaching 0.1 dB away from Shannon's limit. The advantage of these codes used to be that the patents have expired so they are free for use. But since the patents of the Turbo codes have also expired, this is no longer an advantage. A possible disadvantage is high encoding complexity. These codes work to produce convolutional codes derived from quasicyclic block codes of LDPC.

6.6 Theory of pulse systems

The easiest way to explain what is happening with a spread spectrum waveform is to look at a simple square wave. A square wave has a 50% duty cycle and is shown in Figure 6.16. This is a time domain representation of a square wave because it shows the amplitude of the signal with respect to time, which can be observed on an oscilloscope. The frequency domain representation of a square wave shows the amplitude (usually in dB) as a function of frequency, as shown in Figure 6.16. The frequency spectrum of a square wave contains a fundamental frequency,



Time representation of a square wave in the time domain



Frequency spectrum of a square wave in the frequency domain—Mag sinx/x

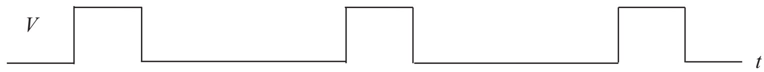
Figure 6.16 Time and frequency domain representations of a pulsed signal with 50% duty cycle

both positive and negative, and the harmonics associated with this fundamental frequency. The negative frequencies do not exist; they are there to help in analyzing the signal when it is upconverted. The fundamental frequency is the frequency of the square wave. If the corners of the square wave were smoothed to form a sine wave, it would be the fundamental frequency in the time domain. The harmonics create the sharp corners of the square wave. The sum of all these frequencies is usually represented by a Fourier series. The Fourier series contains all the frequency components with their associated amplitudes.

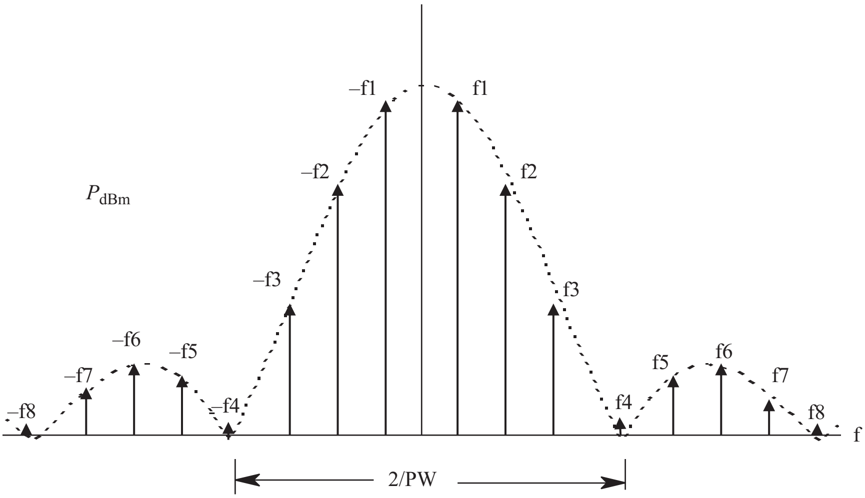
In a square wave, even harmonics are suppressed, and the amplitudes of the odd harmonics form a sinc function, as shown in Figure 6.16. The inverse of the pulse width is where the nulls of the main lobe are located. For a square wave, the nulls are located right at the second harmonic, which is between the fundamental frequency and the third harmonic. Note that all the other nulls occur at the even harmonics, since the even harmonics are suppressed with a square wave (Figure 6.16).

If the pulse stream is not a square wave (50% duty cycle), then, these frequency components move around in frequency. If the pulse width is unchanged and the duty cycle is changed, then only, the frequency components are shifted around and the sinc function is unchanged. Also, the suppressed frequency components are different with different amplitude levels. For example, if the duty cycle is changed to 25%, then, every other even harmonic (4,8,12) is suppressed (Figure 6.17).

Note that not all the even harmonics are not suppressed as they were for the 50% duty cycle pulse because they are not located in the nulls of the sinc pattern.



Time representation of a pulse in the time domain with 25% duty cycle



Frequency spectrum of a pulse in the frequency domain with 25% duty cycle—Mag sinx/x

Figure 6.17 Time and frequency domain representations of a pulsed signal with 25% duty cycle

Also, there are more and more frequencies in the main lobe as the duty cycle gets smaller, and if the pulse width does not change, then, the nulls will always be at the same place, at the inverse of the pulse width. If the pulse width changes, then, the position of the nulls will change.

6.7 PN code

To obtain an intuitive feel for what the spectrum of a PN coded signal is doing on a real-time basis, as the code is changing the pulse widths (variable number of "1"s in a row) and the duty cycle change (variable number of "-1"s in a row). Variations in the pulse widths cause the nulls of the sinc function to change (will not be any larger than 1/chip width). In addition, variations in the duty cycle cause the number of frequency components inside the sinc function to change. This results in a more continuous sinc function spectrum in the frequency domain, with the null of the sinc function being 1/chip width (Figure 6.18).

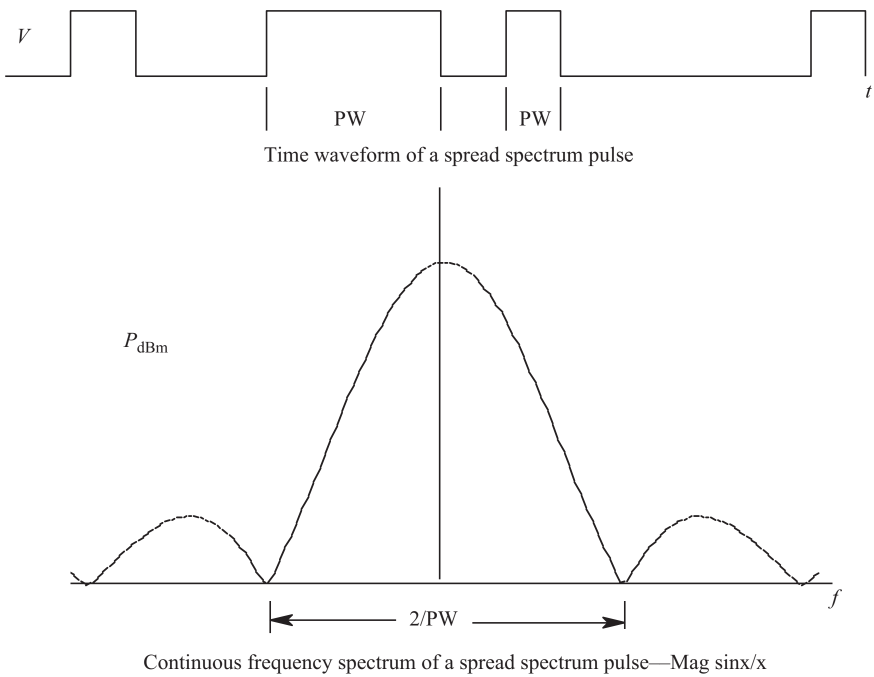


Figure 6.18 Time and frequency domain representation of a spread spectrum continuous waveform

6.8 Summary

A simple approach to understanding probability theory and the Gaussian process is provided to allow the designer to better understand the principles of designing and evaluating digital transmission. Quantization and sampling errors in the analysis and design of digital systems were discussed.

The probability of error and probability of detection are the keys in determining the performance of the receiver. These errors are dependent on the received SNR and the type of modulation used. Three types of error detection are evaluated with the CRC being the best error-detection solution. Error correction is discussed and several approaches are examined to mitigate errors. Systematic linear block codes are used for FEC, and interleaving is a key to improve FEC against burst errors. Theory on pulsed systems, showing time and frequency domain plots, provide knowledge and insight into the design and optimization of digital transceivers.

6.9 Problems

1. Using the fact that the integral of the PDF is equal to 1, what is the percent chance of getting this problem wrong if the chance of getting it right is 37%?
2. What is the expected value of x if $f_x(x) = 0.4$ when $x = 1$ and $f_x(x) = 0.6$ when $x = 2$? What is the mean?
3. Is the answer in problem 2 closer to 1 or 2, and why?
4. What is the $E[x^2]$ in problem 2?
5. What is the variance in problem 2?
6. What is the standard deviation of problem 2?
7. What is the probability the signal will not fall into a 2σ value with a Gaussian distribution?
8. Given a system with too much quantization error, name a design change that can reduce the error?
9. What is the probability of receiving 20 pulses if the probability of detection is 0.98 for one pulse?
10. What is the probability that the error occurs because of one lost pulse in problem 9?
11. What is the difference between probability of error and BER?
12. What are the three basic types of error detection? Which is the best type?
13. What are the two basic types of error correction?
14. How does interleaving maximize the error-correction process against burst errors?

Further reading

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Chapter 7

Multipath

Multipath is a signal transmission path that is different from the desired, or direct, signal transmission path used in communications and radar applications. The amplitude, phase, and angle-of-arrival of the multipath signal interfere with the amplitude, phase, and angle-of-arrival of the desired or direct path signal. This interference can create errors in angle-of-arrival information and in received signal amplitude and phase in data link communications. The amplitude can be larger or smaller depending on whether the multipath signals create constructive or destructive interference. Constructive interference is when the desired signal and the multipath signals are more in phase and add in amplitude. Destructive interference is when the desired signal and the multipath signals are more out of phase and subtract in amplitude.

It is similar to a pool table, where you can hit a ball by aiming directly at the pocket or you can bank it off the table with the correct angle to the pocket. If both balls are in motion, they may interfere with each other at the input to the pocket (Figure 7.1).

The problem with multipath is that the signal takes both paths and interferes with itself at the receiving end. The reflected path has a reflection coefficient that determines the phase and the amplitude of the reflected signal, which is different from the direct path. Also, the reflected path length is longer, which produces a signal with a different amplitude phase. If the phase of the reflected path is different, for example, 180° out of phase from the direct path, and the amplitudes are the same, then the signal is canceled out and the receiver sees very little to no signal.

Fortunately, the reflected path is attenuated, since it is generally a longer path length than the direct path which means it has more free-space attenuation. Also, the level of the multipath depends on the reflection coefficient and the type of multipath. It does not completely cancel out the signal but can alter the amplitude and phase of the direct signal. The effect of multipath can cause large variations in the received signal strength. Consequently, multipath can affect coverage and accuracy where reliable amplitude or phase measurements are required.

Angle-of-arrival errors are called glint errors. Amplitude fluctuations are called scintillation or fading errors. Therefore, the angle-of-arrival, the amplitude, and the phase of the multipath signal are all critical parameters to consider when analyzing the effects of multipath signals in digital communications receivers. Spread spectrum systems contain a degree of immunity from multipath effects since these effects vary with frequency. For example, one frequency component for a given range and angle may have multipath that severely distorts the desired

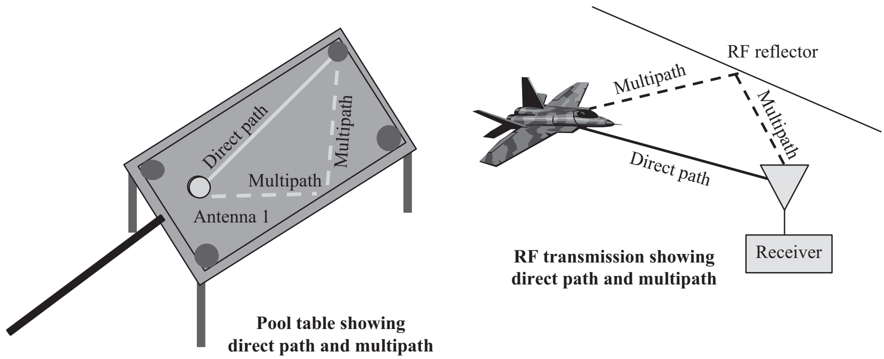

Figure 7.1 Pool table analysis can be used to describe RF signal reflections

signal, whereas another frequency may have little effect. This is mainly due to the difference in the wavelength of the different frequencies. This loss is generally hard to quantize, since there are many variables and many potential paths. Multipath is constantly changing, and certain conditions can adversely affect the coverage and the phase measurement accuracy. Careful positioning or siting of the antennas in a given environment is the most effective way to reduce the effects of multipath. Also, blanking methods to ignore signals received after the desired signal for long multipath returns are often used to help mitigate multipath. And finally, antenna diversity is a very good method to reduce multipath.

7.1 Basic types of multipath

Multipath reflections can be separated into two types of reflections—specular and diffuse—and are generally a combination of these reflections. Specular multipath is a coherent reflection, which means that the phase of the reflected path is relatively constant with relation to the phase of the direct path signal. This type of reflection usually causes the greatest distortion of the direct path signal because most of the signal is reflected toward the receiver. Diffuse multipath reflections are noncoherent with respect to the direct path signal. The diffuse multipath causes less distortion than the specular type of multipath because it reflects less energy toward the receiver and usually has a noise-like response due to the random dispersion of the reflection. Both types of multipath can cause distortion to a system, which increases the error of the received signal and reduces coverage according to the link budget. Multipath effects are included as losses in the link budget (see Chapter 1). Specular reflection is analyzed for both a reflection off a smooth surface and a rough surface.

7.2 Specular reflection on a smooth surface

Specular reflections actually occur over an area of the reflecting surface (which is defined as the first Fresnel zone, similar to the Fresnel zones found in optics).

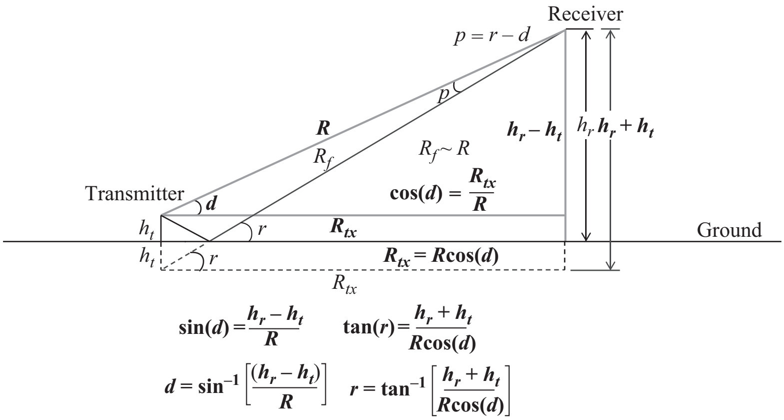

Figure 7.2 Single-ray specular multipath analysis

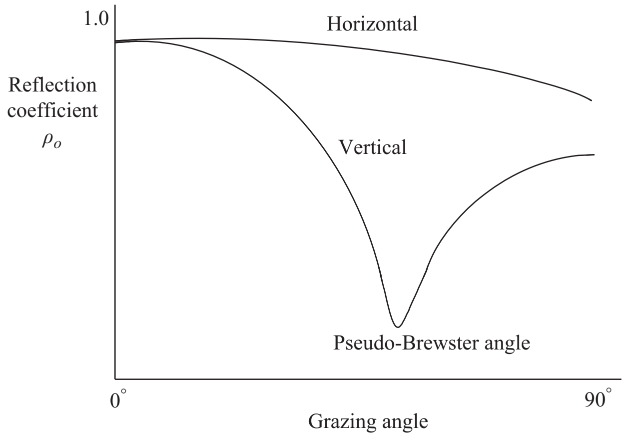

Figure 7.3 Reflection coefficient vs grazing angle for different antenna polarizations

Most of the time, the reflecting area is neglected and geometric rays are used. These rays obey the laws of geometrical optics, where the angle of incidence is equal to the angle of reflection, as shown in Figure 7.2.

For a strictly specular reflection (referred to as a smooth reflection), the reflection coefficient (ρ_o) depends on the grazing angle, the properties of the reflecting surface, and the polarization of the incident radiation. ρ_o assumes a perfectly smooth surface. This reflection coefficient is complex. A smooth reflection has only one path directed toward the receiver, as shown in Figure 7.2.

The magnitude of the "smooth" reflection coefficient (ρ_o) is plotted in Figure 7.3 as a function of the reflection coefficient and the grazing angle, with plots showing both horizontal and vertical polarization effects of the incident radiation.

The graph shows that for vertical polarization near the pseudo-Brewster angle the reflection coefficient is very small, a phenomenon that is also observed in optics. This phenomenon can be used to minimize multipath reflections. For horizontal polarization, the reflection coefficient is fairly constant but starts dropping off as the grazing angle approaches 90°. Not only does the magnitude of the reflected radiation change, but the phase of the reflected radiation is also modified. A phase shift of the incident radiation can vary from near 0° to 180°, depending on the polarization of the incident radiation and the conductivity and dielectric constant of the reflecting surface.

7.3 Specular reflection on a rough surface

Generally the ideal reflecting case must be modified because reflecting surfaces (usually the earth's surface) are not perfectly smooth. The roughness of the reflecting surface decreases the amplitude of the reflection coefficient by scattering energy in directions other than the direction of the receiving antenna. To account for the loss in received energy, a scattering coefficient (ρ_s) is used to modify the smooth reflection coefficient. The smooth reflection coefficient is multiplied by the scattering coefficient, creating an overall modified reflection coefficient to describe a specular surface reflection on a rough surface. ρ_s is the root mean square (RMS) value. This coefficient can be defined as a power scattering coefficient:

$$\overline{\rho}_{\mathrm{s}}^2 = e^{\left[-\left(\frac{4\pi h \sin d}{\lambda}\right)\right]^2}$$

where ρ_s is the power scattering coefficient or reflection coefficient for a rough specular surface, h is the RMS height variation (normally distributed), d is the grazing angle, and λ is the wavelength.

The roughness of the reflecting surface modifies the smooth reflection coefficient and produces a noncoherent, diffuse reflection. If the variations are not too great, then a specular analysis is still possible and ρ_s will be the scattering modifier to the smooth specular coefficient (ρ_o). This analysis can be used if the reflecting surface is smooth compared with a wavelength. The Rayleigh criterion is used to determine if the diffuse multipath can be neglected. It looks at the ratio of the wavelength of the radiation and the height variation of the surface roughness and compares it to the grazing angle or incident angle. The Rayleigh criterion is defined as follows:

$$h_d \sin d < \lambda/8$$

where h_d is the peak variation in the height of the surface, d is the grazing angle, and λ is the wavelength.

If the Rayleigh criterion is met, then the multipath is a specular reflection on a rough surface, and ρ_o and ρ_s are used to determine the multipath. If the Rayleigh criterion is not met, then the diffuse multipath coefficient ρ_d must be taken into account for complete multipath analysis.

7.4 Diffuse reflection

Diffuse multipath has noncoherent reflections and is reflected from all or part of an area known as the *glistening surface* (Figure 7.4). It is also called this because in the optical (visible) equivalent the surface can sparkle when diffuse reflections are present. The boundaries of the glistening surface are given by:

$$y = \pm \frac{X_1 X_2}{X_1 + X_2} \left(\frac{h_r}{X_1} + \frac{h_t}{X_2} \right) \sqrt{{\beta_0}^2 - \frac{1}{4} \left(\frac{h_r}{X_1} - \frac{h_t}{X_2} \right)^2}$$

where β_0 is the reflection angle, h_r is the height of receiver antenna, h_t is the height of transmitter antenna, X_1 is the distance from the transmitter antenna base to the point of reflection, and X_2 is the distance from the receiver antenna base to the point of reflection.

Diffuse reflections have random amplitudes and phases with respect to the amplitude and phase of the direct path. The amplitude variations follow a Rayleigh distribution, while the phase variations are uniformly distributed from 0° to 360°. The phase distribution has zero mean (i.e., diffuse multipath appears as a type of noise to the receiver).

For diffuse reflection, a diffuse scattering coefficient is multiplied by ρ_o to obtain the diffuse reflection coefficient. Calculating the diffuse scattering coefficient (ρ_d) is more tedious than calculating the specular scattering coefficient. The area that scatters the diffuse multipath cannot be neglected, as was done in the specular multipath case. One procedure to determine ρ_d is to break up the glistening area into little squares, each square a reflecting surface, and then calculate a small (δ) diffuse scattering coefficient. The scattering coefficients are summed up and

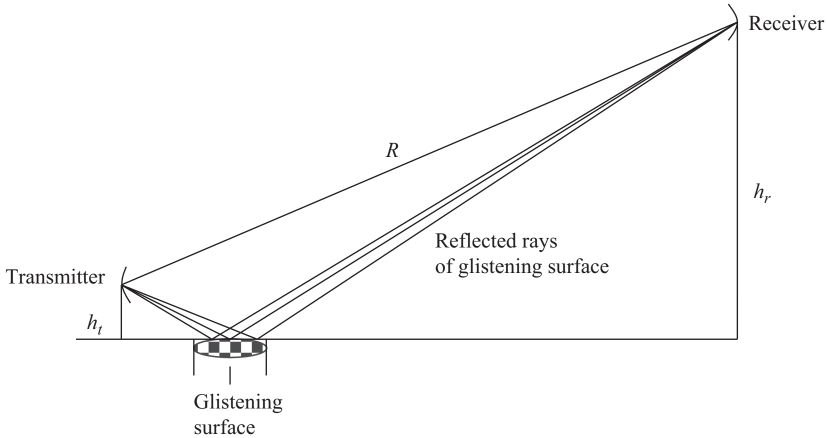

Figure 7.4 Multipath over a glistening surface

the mean is calculated, which gives the overall diffuse scattering coefficient (ρ_d). This can be converted to a power coefficient:

$$\rho_d{}^2 \cong \frac{1}{4\pi\beta_0{}^2}\int \frac{R^2 dS}{X_1{}^2 X_2{}^2} = \frac{1}{2\pi\beta_0{}^2}\int \frac{R^2 y dX}{(R-X)^2 X^2}$$

where R is the range and X is the ground range coordinates.

This is for a low-angle case. This diffuse scattering coefficient is derived based on the assumption that the reflecting surface is sufficiently rough that there is no specular multipath component present. The roughness factor has been defined as F_d^2 and modifies the diffuse scattering coefficient (ρ_d):

$$F_d{}^2 = 1 - \rho_s{}^2$$

A plot of the specular and diffuse scattering coefficients with respect to roughness criteria is shown in Figure 7.5. With a roughness factor greater than 0.12, the reflections become predominantly diffuse. The diffuse reflection coefficient amplitude reaches approximately 0.4, which is 40% of the amplitude of the smooth reflection coefficient, whereas the specular coefficient reaches 1.0, or 100% of the smooth reflection coefficient. Precautions need to be taken when using the plots in Figure 7.5 since ρ_d does not include the roughness factor F_d^2 and may cause significant errors in the analysis of actual values.

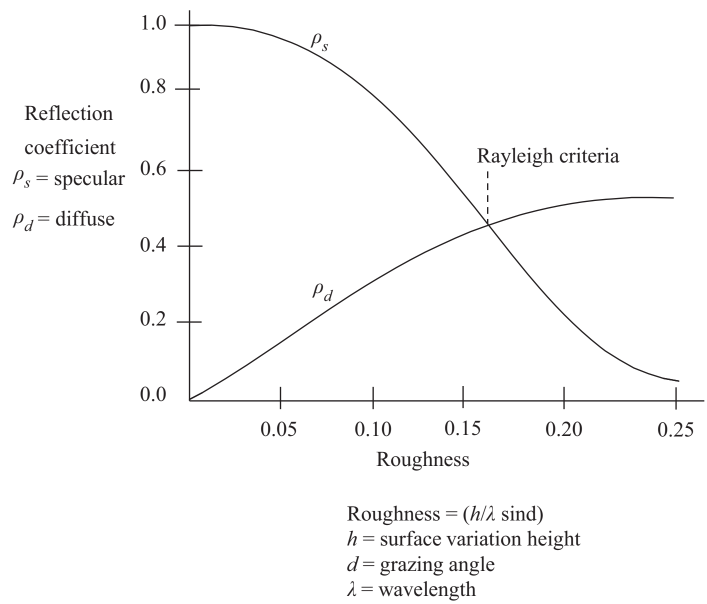

Figure 7.5 Reflection coefficient vs roughness factor

7.5 Curvature of the earth

For most communications systems other than satellite links, the divergence factor D caused by the curvature of the earth can be taken as unity. However, if a particular scenario requires the calculation of the divergence factor, it is equal to:

$$D = \lim_{f-0}\sqrt{\frac{A_r}{A_f}}$$

where A_r is the area projected due to a round earth and A_f is the area projected due to a flat earth.

A diagram showing the different areas that are projected for both the round and flat earth is shown in Figure 7.6. The curvature of the earth produces a wider area of reflection. The ratio of the areas is used to calculate the divergence factor. Note that for most systems, the area difference is very small and generally can be neglected.

7.6 Pulse systems (radar)

Multipath can affect radar systems by interfering with the desired signal. Leading edge tracking can be used to eliminate a large portion of the distortion caused by multipath because the multipath arrives later in time than the direct path. In leading edge tracking, only the first portion of the pulse is processed, and the rest of the pulse, which is distorted due to multipath, is ignored. If the grazing angle is very small, then the time of arrival (TOA) for both direct path and multipath signals is about the same. For this case, the angle error between the direct path and multipath is very small. If the multipath is diffuse, then the multipath signal will appear as an

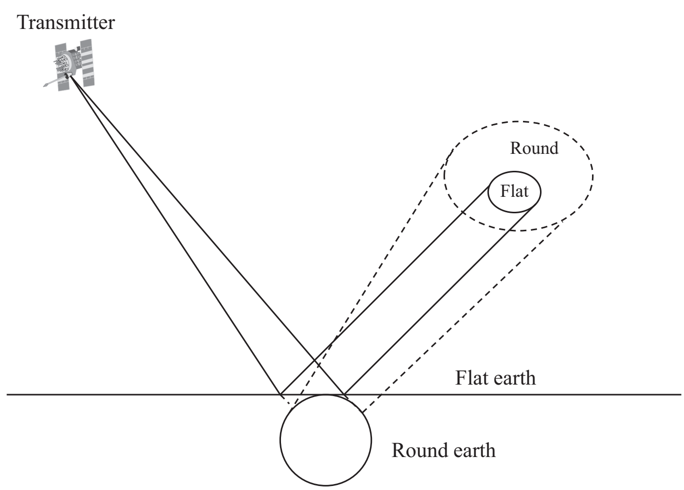

Figure 7.6 Different areas for both the round and flat earth

increase in the noise floor of the direct path and a slight angle variation will occur due to the multipath mean value coming at a slightly different angle. If the multipath is specular, then the multipath signal will increase or decrease the actual amplitude of the direct path signal and also create a slight angle error. If one or more of the antennas are moving, then the specular reflection will also be a changing variable with respect to the phase of the multipath signal and the direct path signal, thus producing a changing amplitude at the receiver. As the grazing angle increases, the angle error also increases. However, the TOA increases, which means that leading edge tracking becomes more effective with larger grazing angles.

One other consideration is that the pulse repetition frequency (PRF) of the radar must be low enough so that the time delay of the multipath is shorter than the time between the pulses transmitted. This prevents the multipath return from interfering with the next transmitted pulse.

Another consideration is the scenario where a surface such as a building or other smooth surface gives rise to specular reflection with a high reflection coefficient where the multipath could become fairly large. For a stationary situation, this is a real problem. However, if one of the antennas is moving, then the multipath is hindered, depending on the velocity of the antenna and the area of the reflector, for only the time the angle of reflection is right. The processor for the receiver could do an averaging of the signal over time and eliminate some of these problems.

7.7 Vector analysis approach

Different approaches can be taken to calculate the resultant effects of the reflected energy. One method is vector addition. The reference vector is the direct path vector with given amplitude and a zero reference angle. The coherent reflection vector, C, for specular reflection is now calculated with the phase referenced to the direct path vector:

$$\overline{C} = \overline{D}\rho_0\rho_s e^{j2\pi\frac{dR}{\lambda}}$$

where D is the divergence factor, ρ_o is the specular reflection coefficient, ρ_s is the scattering coefficient, dR is the path length difference, and λ is the wavelength.

Determining an accurate vector for ρ_s is difficult. Note that ρ_s represents an average amplitude distributed value that is a function of the distribution of the surface.

The noncoherent vector (I) for diffuse reflection is calculated as:

$$\overline{I} = \frac{\overline{D}\rho_0\overline{\rho_d}}{\sqrt{2}}(I_1 + jI_2)e^{j2\pi\frac{dR}{\lambda}}$$

where dR is the path length difference, D is the divergence factor, and $I_1 + I_2$ is the independent zero mean unit variance normalized Gaussian processes.

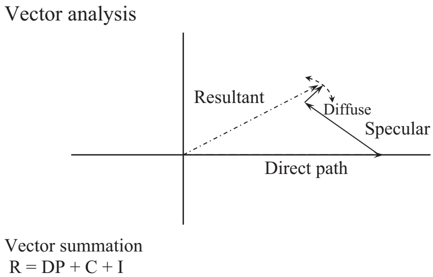

Figure 7.7 Different approaches to analyzing multipath

The specular vector C and the diffuse vector I are summed with the reference or direct path vector DP to produce the resultant vector R, as shown in Figure 7.7.

Designing an accurate model for ρ_d is difficult. Observed test data from known systems can aid in the selection and verification of the scattering coefficients ρ_s and ρ_d.

Summing the power vectors for determining the error caused by reflections is to sum the powers of each type of reflection with the direct power giving the resultant power:

$$\overline{P}_R = \overline{P}_{DP} + \overline{P}_C + \overline{P}_I$$

where P_{DP} is the direct power, P_C is the mean specular power, and P_I is the mean diffuse power.

Mean power reflection coefficients are used. The resultant mean specular power is:

$$\overline{P}_C = \overline{P}_{DP}\rho_0{}^2\rho_s{}^2$$

The mean diffuse power is equal to:

$$\overline{P}_I = \overline{P}_{DP}\rho_0{}^2\rho_d{}^2$$

A divergence factor can be included to account for the curvature of the earth.

7.8 Multipath mitigation techniques

Some of the techniques used to reduce and mitigate multipath are the following:

- Leading edge tracking for radars. The multipath is delayed with respect to the direct path; therefore, the leading edge of the waveform contains the signal and not the multipath. This assumes a low PRF to prevent long multipath from affecting the next radar pulse. For long multipath, the signal amplitude is smaller which will reduce the interference to the desired signal.

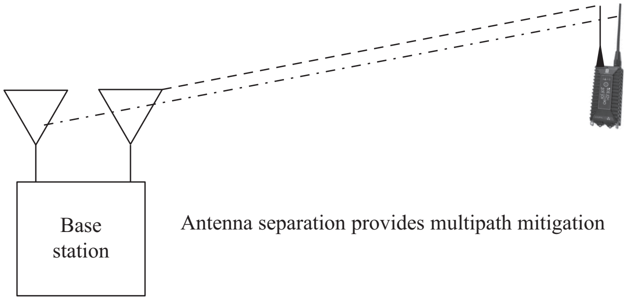

Figure 7.8 Antenna diversity to help mitigate multipath

- Movement changes the specular multipath to provide an average error. Multipath is position sensitive, so movement will change the multipath nulls.
- Antenna design to prevent low-angle multipath. This technique is used for global positioning systems, where the desire is to have the mask angle or angle to the earth low but at the same time prevent multipath signals that are reflecting off the earth.
- Spread spectrum. Nulls caused by multipath are frequency dependent, so with a spread spectrum system spread across a wide band of frequencies, the multipath only affects a portion of the spectrum.
- Antenna diversity. Since multipath is dependent on position, antennas can be located at different positions so that as one antenna is in a multipath null the other antenna is not.

7.8.1 Antenna diversity

Antenna diversity is a technique to help mitigate multipath. It requires the data link to have two or more antennas at the receiver. The signals in these different paths are either selected to provide the best signal that is not interfered with by multipath, or combine both of the inputs from the antennas. The theory is that when one antenna is in a multipath null, the other is not. This works very well with only two antennas with the antennas separated a short distance to prevent both antennas from being in the null (see Figure 7.8). This technique is used in many applications in wireless communications.

7.9 Summary

Multipath affects the desired signal by distorting both the phase and the amplitude. This can result in a lost signal or a distortion in the TOA of the desired signal. Multipath is divided into two categories: specular and diffuse. Specular multipath generally affects the system the most, resulting in more errors. Diffuse multipath is more noise-like and is generally much lower in power. The Rayleigh criterion is used to determine if the diffuse multipath needs to be included in the analysis. The curvature of the earth can affect the analysis for very long-distance multipath.

One of the ways to reduce the effects of multipath is to use leading edge tracking so that most of the multipath is ignored. Some approaches for determining multipath effects include vector analysis and power summation. Several methods of multipath mitigation were discussed, including using multiple antennas for antenna diversity.

7.10 Problems

1. What is the difference between glint errors and scintillation errors?
2. Which type of multipath affects the solution the most? Why?
3. What is the effect of multipath on an incoming signal if the signal is vertically polarized at the pseudo-Brewster angle?
4. What is the effect of multipath on an incoming signal if the signal is horizontally polarized at the pseudo-Brewster angle?
5. What is the criterion for determining which type of multipath is present?
6. According to the Rayleigh criterion, what is the minimum frequency at which the multipath will still be considered specular for a peak height variation of 10 m and a grazing angle of 10°?
7. What is the divergence factor, and how does it affect the multipath analysis?
8. How do most radars minimize multipath effects on the radar pulses?
9. Graphically show the resultant vector for a reference signal vector with a magnitude of 3 at an angle of 10° and the smooth specular multipath signal vector with a reflection coefficient of 0.5 at an angle of 180° with respect to the signal vector.
10. What is the main difference between the vector summation approach to multipath analysis and the power summation approach? How does this approach affect the reflection coefficient?
11. How does antenna diversity help mitigate multipath?

Further reading

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Chapter 8

Improving the system against jammers

The receiver is open to reception of not only the desired signal but also all interfering signals within the receiver's bandwidth, which can prevent the receiver from processing the desired signal properly (Figure 8.1). Therefore, it is crucial for the receiver to have the ability to eliminate or reduce the effects of the interfering signals or jammers on the desired signal. These are divided into two groups, cosite and friendly jammers, which are often referred to as interferers, and unfriendly jammers that purposely jam the signal. The focus of this chapter is to address the types of unfriendly jammers and ways to mitigate their effect on the desired communication signal. Here is a list of several unfriendly jammers that are currently used in the environment:

- Spot jammer—jams a single frequency
- Barrage—sends out a barrage of frequencies
- Continuous wave (CW)—high power at frequency of interest
  - Receiver only has so much process gain
  - Saturates or desensitizes the receiver
- Repeater jammer—repeats the waveform with delay
- Cognitive—adapts to the receiver's antijam techniques
- Burst or pulsed—high-power pulse, low average power which captures the automatic gain control (AGC)

8.1 Unfriendly jammers

There are several reasons that unfriendly jammers are used against communications. The military, as part of strategic planning, may decide that it is advantageous to develop a jammer to impede the enemy's communication link. These jammers, along with methods to mitigate the effects of these jammers, are the discussion that follows.

8.1.1 Spot jammer

Spot jammers are focused on jamming a specific frequency. Their advantages are as follows:

- Good if the frequency of operation is known
- Good if the frequency of operation does not change
- All the power can be focused at one frequency

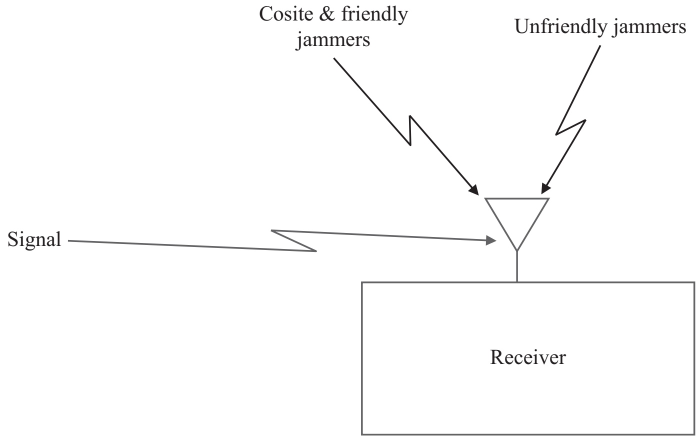


Figure 8.1 The receiver accepts both the desired signal and jammers

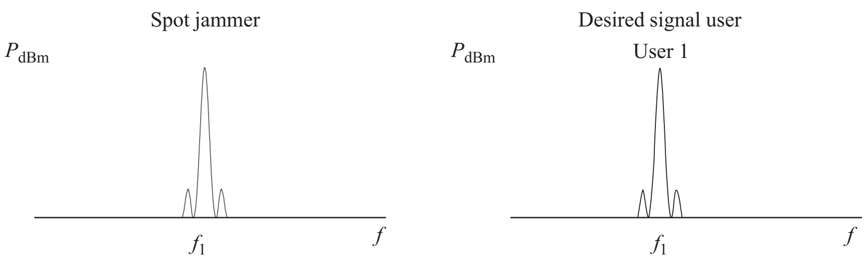


Figure 8.2 Spot jammer affects a signal frequency

This is a very effective jammer at one frequency, and the jammer can focus all of the power as a narrowband jammer at that frequency (Figure 8.2). However, a frequency diverse signal that has the ability to change frequency, or an agile frequency hopping signal is tolerant to this type of jammer. Also, spread spectrum communications contain process gain to reduce the effects of a narrowband jammer. Another challenge is the jammer needs to know what the frequency is or search for it.

8.1.2 Barrage jammer

A barrage jammer is intended to jam a broadband spectrum by sending a barrage of frequencies so that frequency diverse signals are vulnerable. The advantages of a barrage jammer are as follows:

- Covers a wideband
- Jams multiple users on different frequencies
- Low complexity, easy to implement
- Does not require a priori knowledge of the frequency of the signal

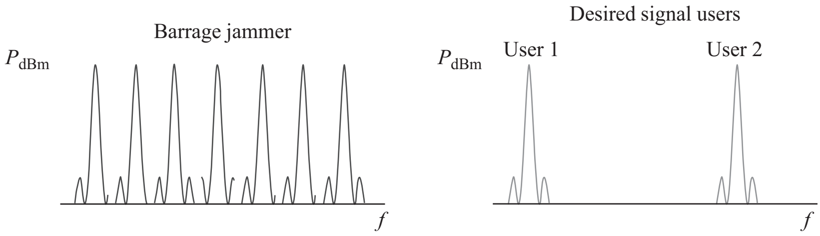


Figure 8.3 Barrage jammers affect multiple users and frequencies

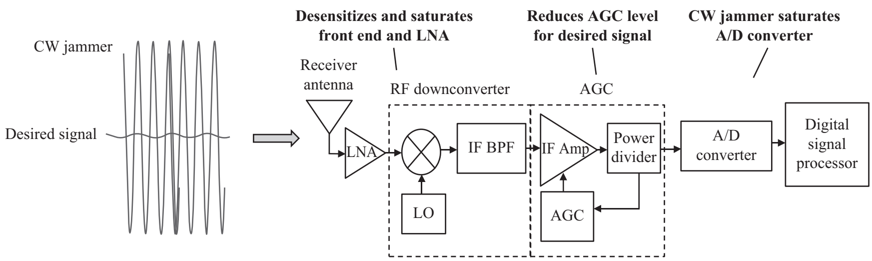


Figure 8.4 CW jammer effects on the desired signals independently of frequency

This type of jammer is effective because it jams the whole band of interest and multiple users on different frequencies (Figure 8.3). However, the jammer requires a large amount of average power in order to cover a broadband spectrum. Therefore, there is lower jamming power for each specific frequency being jammed.

8.1.3 Continuous wave (CW) jammer

The CW jammer is basically the same as the spot jammer but generally has more power and is only a CW and does not have any modulation. This is a very high-power jammer on a single frequency. The advantages of a CW jammer are as follows:

- High-power output
- Jams the signal at the given frequency
- Jams other waveforms at different frequencies—overcomes receiver isolation
- Saturates the receiver front end—desensitizes the receiver at all frequencies

The CW jammer is effective against a known frequency signal and also due to high power can jam others nearby on different frequencies, since the high-power CW jammer can overcome the receiver's isolation and also saturate or desensitize the receiver front end (Figure 8.4). However, this requires very high average power output which has high cost and size. In addition, good receiver filtering can help overcome the CW jammer. Furthermore, this jammer can be detected at the signal source for electronic countermeasures against the jammer.

8.1.4 Repeater jammer

This type of jammer is used for frequency agile signals which have the ability to move to different frequencies or a frequency hopping system. The advantages of the repeater jammer are as follows:

- Uses the input signal to determine the jammer's frequency
- Spoofs the receiver by sending a delayed version of the desired signal
- Adapts to the changing input signal
- Simple to implement

A repeater jammer is powerful in that it tries to repeat the incoming signal with the delay of the response time to process the data and send it back out as a delayed jamming signal (Figure 8.5). For a fast-changing signal, this jammer may have a failure-to-follow condition where it can't keep up with the changing input signal. In addition, it is not very effective against fast frequency hopping or spread spectrum signals and is not efficient against multiple signals.

8.1.5 Cognitive jammer

The cognitive jammer is considered the biggest threat to communication signals. It has the ability to assess the environment, monitor the desired data link, and then create and send out the optimal jammer for the signals that are present in the environment (Figure 8.6). The advantages of this type of jammer are as follows:

- Ability to adapt to the optimal solution
- Monitor the jammers and use learning and reasoning to develop the best jammer to use
- Monitors and determines the best pulse duration
- Determines and adapts to the amount of power required

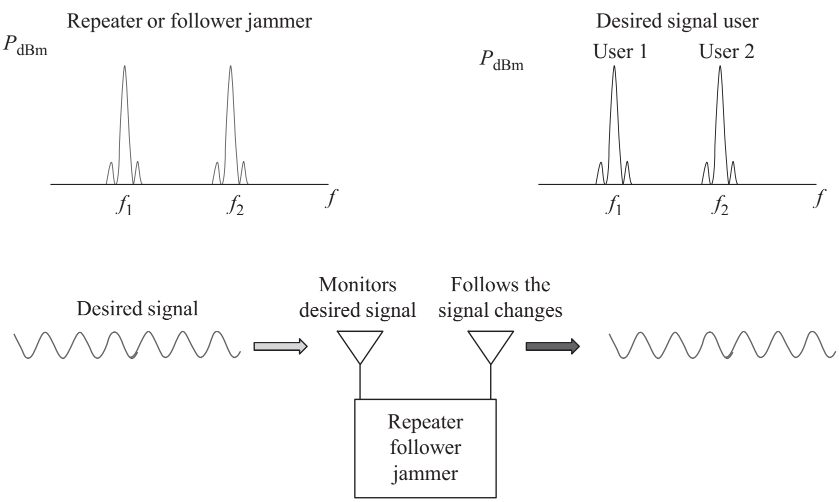


Figure 8.5 Repeater or follower jammer follows the changes of the desired signal

- Provides agility to change as the receiver incorporates antijam solutions
- Senses the changes and adapts to optimum jamming
- Learns from the past receiver's mode of operation and makes more intelligent changes in the future

The cognitive jammer is a smart, adaptable, versatile jammer and is constantly improving the jammer effects on the data link. Complexity and time-to-change are some of the disadvantages of this type of jammer, but overall, this can become the optimal jammer as they are being developed.

8.1.6 Burst or pulsed jammer

A burst or pulsed jammer is used extensively to minimize the average power and is very effecting against AGC receivers (Figure 8.7). The advantages of a burst or pulse jammers are as follows:

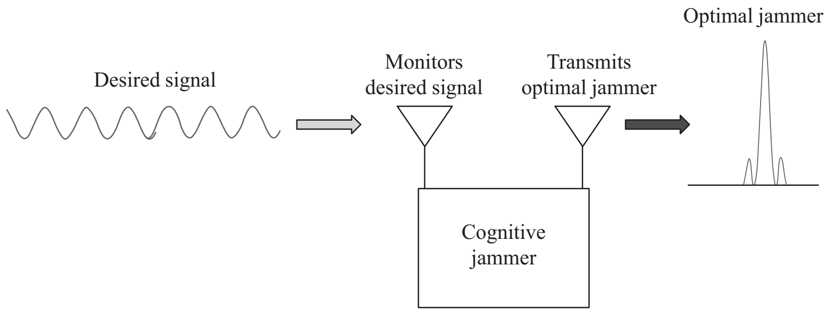


Figure 8.6 Cognitive jammer monitors the signal and optimizes the type of jammer

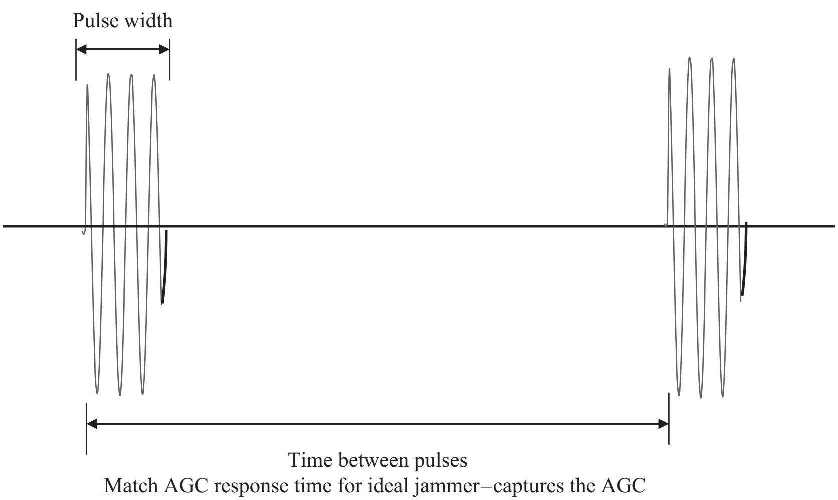


Figure 8.7 An example of a burst jammer

- High-power output
- Jams for the pulse duration
- Recharges for the next high-level jamming pulse
- Low average power, small size
- Captures the AGC for maximum effectiveness

This type of jammer is one of the best jammers for direct sequence spread spectrum modulation signals. A burst jammer is generally a high-amplitude narrowband signal relative to the desired signal and is present for a short period of time. The disadvantage of this type of jammer is that it only jams the signal during the pulse duration. In addition, spread spectrum systems can overcome the errors during the burst, and continuous signals use forward error correction and interleaving to correct burst/pulse errors. Typical bursts range from 0 to 40 dB signal-to-jammer ratio (SJR), with a duration of 0.5 to 1,000 ms. The burst jammer affects the receiver as follows:

- The high amplitude of the burst saturates the AGC amplifiers, detectors, and the processor. The information is lost during the burst time and also during the recovery time of each of the devices.

The burst can capture the AGC.

8.1.7 Capturing the AGC

All AGC receivers are using feedback to change the power level in order to increase the dynamic range. If the burst/pulse jammer knows the recovery time of the AGC, then it can optimize the duty cycle of the jammer to coincide with the AGC time constant (Figure 8.7). The AGC circuit in the receiver senses an increase in power level and reduces the gain in the receiver. This eliminates the ability of the receiver to process small-level signals. When the burst is gone, then the AGC increases the gain in the receiver for low-level signals. This takes time due to the design of the AGC. Once it is back to receiving the low-level signals, the jammer sends out another burst (Figure 8.8). The jammer has captured the AGC and locks the burst jammer's duty cycle to the AGC's response time. The burst or pulsed jammer tries to capture the AGC or finds out the recovery time of the AGC in order to maximize the effectiveness of the jammer.

This optimizes the effectiveness of the jammer signal.

8.2 Techniques to reduce jammers

There are several techniques to reduce the jammer effects on a data link. One of the ways is to use antenna siting or mounting the antenna pointing away from jammer or structures that cause reflections from the jammer. This is possible for a stationary system where both the signal and jammer are not moving. This is accomplished by raising or lowering the antenna, or mounting it in a different location where the receiver is less susceptible to the jammer effects. An example of this is a GPS antenna receiving the 13th harmonic of a very high frequency (VHF) radio on board a Navy ship.

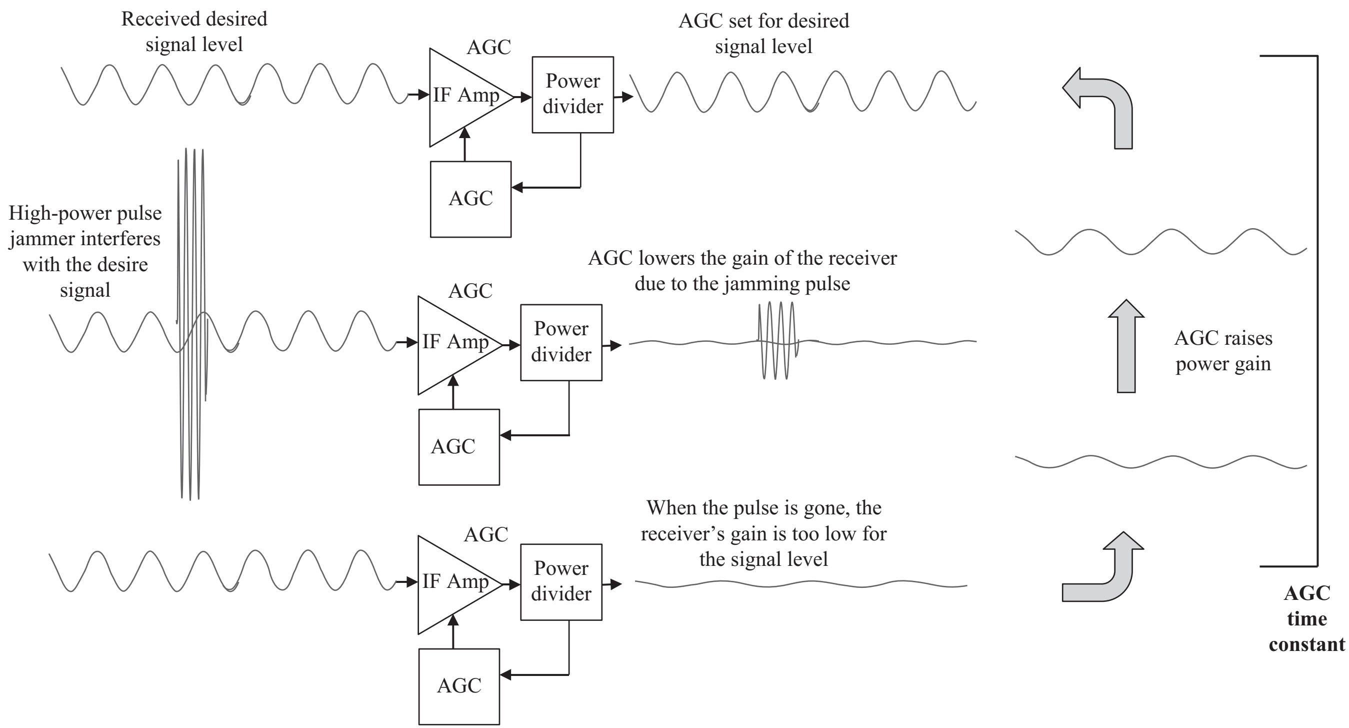


Figure 8.8   Capturing the AGC

The GPS patch antenna was mounted higher up on the ship so that the VHF did not jam the low-level GPS signal. The antenna design is another way to mitigate jammers. An antenna can be designed to point away from the jammer, steer a null where the jammer is located, or raise a mask angle to prevent ground multipath and low-angle jammers. These techniques will be further discussed in the cognitive part of the book. For burst jammers, a way to reduce the effects of the jammer is to use a burst clamp.

8.2.1 Burst clamp

One method of reducing the effects of a burst jammer is to use a burst clamp, which detects the increase in power at radio frequency (RF) and prevents the burst from entering the receiver. To determine the threshold for the power detector, the previous AGC voltage is used. The process gain and the bit error rate (BER) are also factors in determining the desired threshold. The threshold durations are determined by the noise spikes for a minimum and the expected changes in signal amplitude for a maximum. The burst clamp performs four functions during a received burst jammer, which are as follows:

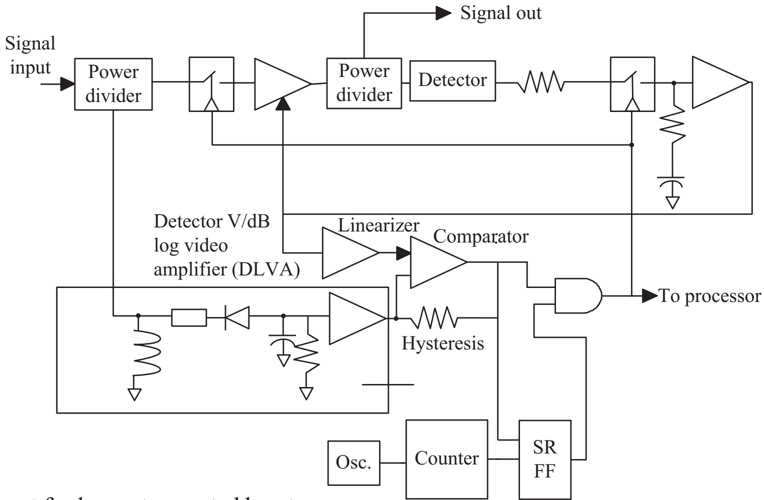
- Detects the power level to determine the presence of a burst.
- Switches out the input signal to the receiver and the AGC amplifier.
- Holds the AGC voltage to the previous value before the burst occurred.
- Informs the decoder that a burst is present.

When the burst is off, the burst clamp performs the following four functions:

- Detects the power level to determine if the burst is gone
- Switches the input signal back into the receiver and the AGC amplifier
- Informs the decoder that a burst is gone
- Restart the detection over a time frame to prevent lockup

A detector log amplifier is used to determine the power level of the burst (Figure 8.9). The log amplifier compensates for the detector and gives a linear response of power in dB to volts ratio output. The detected output is compared with the AGC voltage plus a threshold voltage to determine whether or not a burst is present. If a burst is present, the counter in the time-out circuit is enabled. This switches out both the intermediate frequency (IF) path and the AGC path, prevents the burst from continuing through the circuit, and holds the AGC voltage at the level set before the burst occurred. The counter time-out is set for the longest expected burst and then resets the flip-flop that closes the switches to allow the AGC voltage to build up again. This prevents the burst clamp from locking up if there is a large change in signal level or when the transmitter is turned on. The speed of the circuitry is important so that the receiver will respond to quick bursts and not allow the AGC voltage to change due to a burst. The response of the IF amplifier is slow enough so that the detection circuitry has time to respond. Some considerations when designing the circuitry are as follows:

- Linearizing and matching over temperature.
- Response time of the burst clamp.



Counter set for longest expected burst
Counter resets FF after time out to close the switches again
SR FF goes high for one count cycle to allow capacitor to charge up

Figure 8.9 Burst clamp used for burst jammers

- False triggering on noise.
- Dynamic range of the detector.
- Detector amplitude dependent on frequency.
- Burst clamp saturation.
- Holding AGC voltage for long burst duration.

The response time needs to be fast enough so that the burst is not in the system longer than the error correction used in the system. The instantaneous dynamic range (IDR) of the system (amplitude) affects the soft decision (various thresholds are used in soft decisions), which can affect the BER regardless of the process gain.

Another method for making a burst clamp is by using a slope detector, as shown in Figure 8.10. This design uses a differentiator to detect the presence of a burst. The window comparator looks at both the rising slope and falling slope of the burst. The rest of the circuitry is basically the same as before. The advantages of this type of clamp are as follows:

- Same threshold for all signal levels.
- No need for curve matching or linearizing.
- Insensitive to slow changes in noise level and signal level.

The disadvantages are as follows:

- Hard to detect the slope (detection error).
- Will not detect slow rise time bursts.
- Difficult to select the proper time constant.
- Slower detection process.

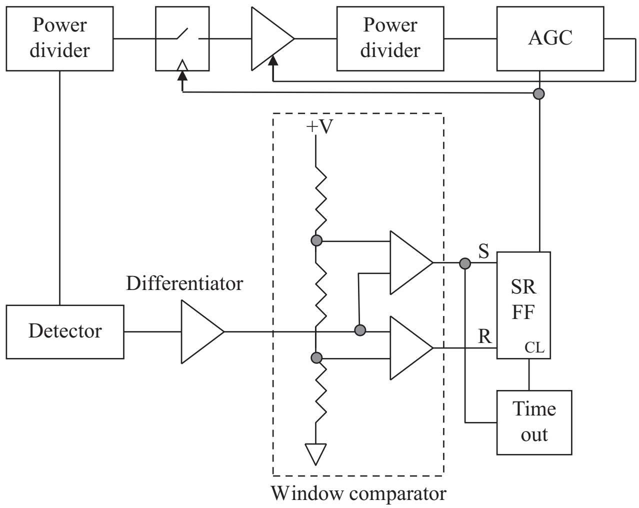


Figure 8.10 Slope detection burst clamp

The placement of the burst clamp depends on the type of receiver used in the system. Most systems use a burst clamp at the IF and possibly one at RF. If the system has an AGC at the RF, then a burst clamp would be needed there also. The receiver either can have one detector at RF and use the sum of the RF and IF AGC voltages for the threshold or can use a detector at both places in the system. The latter would increase the sensitivity of the overall burst detection, depending on the receiver. Most detectors have a sensitivity of around −85 dBm.

8.2.2 Adaptive filter

An adaptive notch filter that minimizes the effect of a narrowband jammer on data links can improve the transceiver. Most adaptive tapped delay line filters function only in baseband systems with bandwidths on the order of a few megahertz. To operate this filter over a wide bandwidth and at a high frequency, several problems must be overcome. Phase delays must be compensated for, and the quadrature channels must be balanced. Therefore, the actual performance of the filter is limited by the ability to make accurate phase delay measurements at these high frequencies.

In situations where the desired signal is broadband, there is a requirement to design a filtering system that can reduce narrowband jamming signals across a very wideband at a high center frequency without an extensive amount of hardware. The effect of the narrowband signals across a very large band can be reduced in several ways.

One such method is the use of spread spectrum techniques to obtain a process gain, which improves the SJR. Process gain is the ratio of the spread bandwidth to

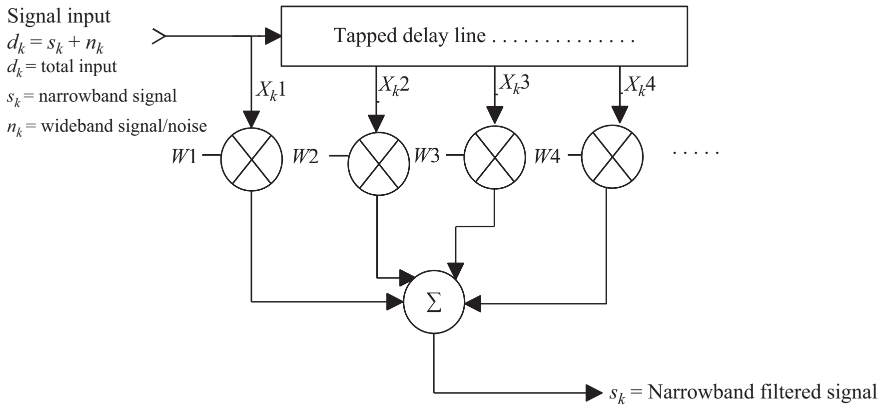


Figure 8.11 Digital finite impulse response (FIR) filter

the rate of information sent. However, there are limitations on the usable bandwidth and on how slow the information can be sent in a given system. Coding techniques and interleaving can improve the system and reduce the effect of the narrowband jammer. However, the amount and complexity of the hardware required to achieve the necessary reduction limits the amount of coding that is, practical in a system. If the bandwidth is already wide, spreading the signal may be impractical, and the spread spectrum techniques will be of no value.

Passive notch filters placed in-line of the receiver can be used to reduce the unwanted signals, but with a large degradation in the wideband signal at the location of the notches. Also, prior knowledge of the frequency of each interferer is required, and a notch for each undesired signal needs to be implemented. Since these notch filters are placed in series with the rest of the system, the overall group delay of a communication link will be altered.

Adaptive filters have been used for noise cancelation using a tapped delay line approach. The noise cancelation filter uses a separate reference input for the noise and uses the narrowband output for the desired signal. In this application, the desired signal is the wideband output, and the reference input and the signal input are the same. The reference signal goes through a decorrelation delay that decorrelates the wideband component, but the narrowband signal, because of its periodicity, remains correlated. When an adaptive filter is configured in this manner, it is called an adaptive line enhancer (ALE).

Adaptive filters, configured as ALEs, have been used to reduce unwanted signals, but they have been limited to relatively narrowband systems and at low frequencies. Adaptive filters can be used in broadband systems with a high carrier frequency, provided certain modifications are made to the system.

8.2.2.1 Adaptive digital filter intuitive analysis

A finite impulse response (FIR) digital filter is shown in Figure 8.11. The signal enters the tapped delay line. Each output of the taps is multiplied by a weight value,

and then they are all summed together. One way to look at this is that each tap is moving up and down according to the input signal and time. Therefore, the sum is equal to a point on the sine wave in time. At another point in time, the signal levels are different in the taps. However, the multiplication and resulting sum equal another point on the sine wave. This continues to happen until the output is a sine wave or the input is a sine wave with a delay. Note that other frequencies will not add up correctly with the given coefficients and will be attenuated.

An adaptive filter uses feedback to adjust the weights to the correct value for the input sine wave compared with the random signal, and in ALE, the sine wave is subtracted from the input signal, resulting in just the wideband signal output (Figure 8.12). Note that there is a decorrelation delay in the ALE that slides the wideband signal in such a way that the autocorrelation of the wideband signal is small with a delayed version of itself. The narrowband signal has high autocorrelation with a delayed version of itself. This technique is used to reduce narrowband jamming in a wideband spread spectrum system. The wideband signal is not correlated with the delayed wideband signal, especially for long pseudonoise (PN) codes. If a long PN code is used, then a delay greater than one chip of the code is not correlated. For longer codes, the correlation is less for a delay greater than a chip (Figure 8.13). The correlator multiplies the code with the delayed version of the code, and the results are integrated. The integration value approaches zero as the code becomes longer.

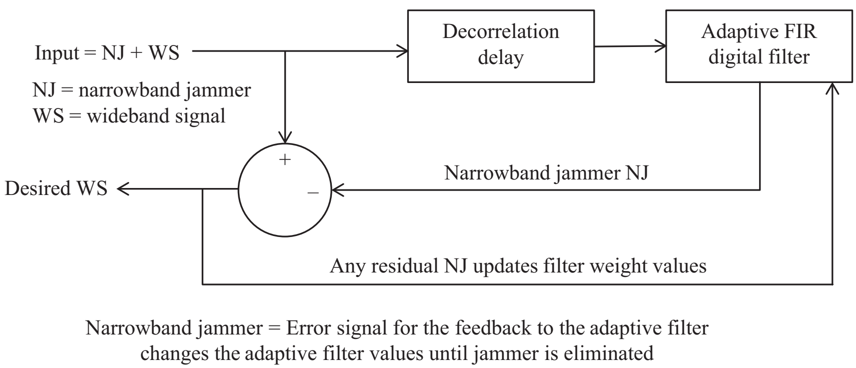


Figure 8.12 Adaptive filter

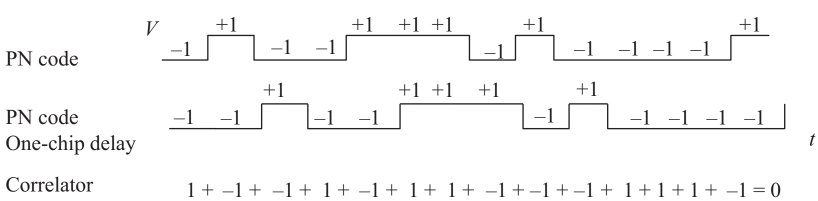


Figure 8.13 Correlation of a wideband PN code with a one-chip delay

8.2.2.2 Basic adaptive filter

A block diagram of a basic adaptive filter configured as an ALE is shown in Figure 8.14. The wideband input (d_k) consists of a narrowband signal (s_k) plus a wideband signal or noise (n_k). The composite signal is split, and one channel is fed to a decorrelation delay indicated by Z^{-D}, which decorrelates the wideband signal or noise. The other goes to a summing junction. The output of the decorrelation delay is delivered to a chain of delays, and the output of each delay, $X_k(i)$, is multiplied by its respective weight values, $W_k(n)$. The weights are drawn with arrows to indicate that they are varied in accordance with the error feedback (e_k). The outputs of all the weights are summed together and produce the estimated narrowband spectral line (y_k). This narrowband signal is subtracted from the wideband plus narrowband input (d_k) to produce the desired wideband signal output, which is also the error (e_k). The name ALE indicates that the desired output is the narrowband signal (y_k). However, for use in narrow-band signal suppression, the wideband or error signal (e_k) is the desired output.

ALEs are different from fixed digital filters because they can adjust their own impulse response. They have the ability to change their own weight coefficients automatically using error feedback, with no a priori knowledge of the signal or noise. Because of this ability, the ALE is a prime choice in jammer suppression applications where the exact frequency is unknown. The ALE generates and subtracts the narrowband interferer, leaving little distortion to the wideband signal. Also, single ALE can reduce more than one narrowband interference signal at a time in a given bandwidth. The adaptive filter converges on the superposition of the multiple narrowband jammers. The adaptive filter scheme has the ability to adapt in frequency and amplitude to the interferer in a specified bandwidth, so exact knowledge of the interferer is not necessary. If the interferer changes in frequency

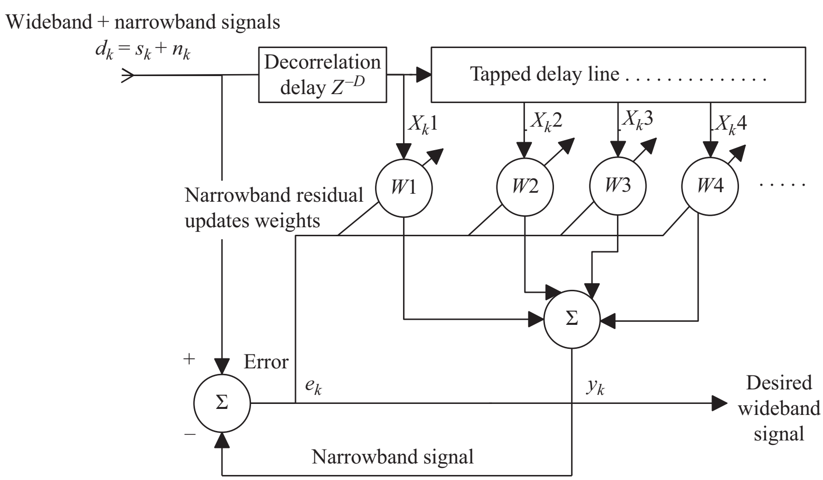


Figure 8.14 Basic adaptive filter configured as an ALE

in the given band, the filter will automatically track the change and reduce the narrowband signal.

8.2.2.3 Least mean square algorithm

The adaptive filter works on the principle of minimizing the mean square error (MSE), $E\left[e_k^2\right]$, using the least mean square (LMS) algorithm. The input to the filter is:

$$d_k = s_k + n_k \text{ (scalar)}$$

where s_k is the narrowband signal (scalar), and n_k is the noise or wideband signal (scalar).

The ALE is designed to minimize the MSE:

$$\begin{aligned}\text{MSE} &= E\left[e_k^2\right]\\ &= E\left[\left(d_k - W_k^T X_k\right)^2\right]\\ &= E\left[d_k^2 - 2d_k W_k^T X_k + \left(W_k^T X_k\right)^2\right]\\ &= E\left[d_k^2\right] - 2E\left[d_k X_k^T\right] W_k + E\left[W_k^T X_k X_k^T W_k\right]\end{aligned}$$

where e_k is the error signal (scalar), W_k^T is the weight values transposed (vector), and X_k is the tap values (vector).

When substituting the autocorrelation and crosscorrelation functions, the definitions give:

$$\text{MSE} = E\left[d_k^2\right] - 2r_{xd} W_k^T + W_k R_{xx}^T W_k$$

where $R_{xx} = E\left[X_k X_k^T\right] =$ autocorrelation matrix, $r_{xd} = E\left[d_k X_k^T\right] =$ crosscorrelation matrix, and $X_k^T W_k = X_k W_k^T$.

To minimize the MSE, the gradient (∇_w) is found and set equal to zero:

$$\nabla_w E\left[e_k^2\right] = 2R_{xx} W_k - 2r_{xd} = 0$$

Solving for W, which is the optimal weight value, gives:

$$W_{\text{opt}} = R_{xx}^{-1} r_{xd}$$

The weight equation for the next weight value is:

$$W_{k+1} = W_k - \mu \nabla_w E\left[e_k^2\right]$$

In the LMS algorithm, e_k^2 itself is the estimate of the MSE. Because the input signal is assumed to be ergodic, the expected value of the squared error can be estimated by a time average. The weight equation then becomes:

$$W_{k+1} = W_k - \mu \nabla_w\left(e_k^2\right) = W_k + 2\mu e_k X_k$$

This is known as the LMS algorithm and is used to update the weights or the filter response. The new weight value, W_{k+1}, is produced by summing the previous

weight value to the product of the error (e_k) times the tap value (X_k) times a scale factor (μ), the latter of which determines the convergence rate and stability of the filter. It has been shown that under the assumption of ergodicity, the weights converge to the optimal solution W_{opt}.

8.2.2.4 Analog/digital adaptive filter

In digital adaptive filters, it is assumed that the time for the error signal to be generated and processed to update the weight values is less than a clock cycle delay of the digital filter clock. However, since part of the feedback loop is analog, delay compensation is required. A quadrature method is used in the frequency conversion processes to allow ease of tuning across a very wideband of operation. This also provides a wider instantaneous bandwidth. Unless the signal can be digitized at the RF, an analog/digital combination needs to be implemented. When ALEs are used at high frequencies, the RF signal must be downconverted using a local oscillator (LO) and a mixer before it can be processed and digitized by the digital filter (Figure 8.15).

A bandpass filter provides a coarse selection of the band for processing. The actual processing bandwidth of the signal is limited to the clock frequency of the digital filter due to the Nyquist criteria to prevent aliasing. After the narrowband signal is generated by the digital filter, it is upconverted, using a mixer and the same LO, to the RF. The signal is fed through a bandpass filter to eliminate the unwanted sideband produced in the upconversion process and eventually is subtracted from the composite signal. The final output is split and used for the error feedback signal, which is downconverted in the same manner as the reference signal.

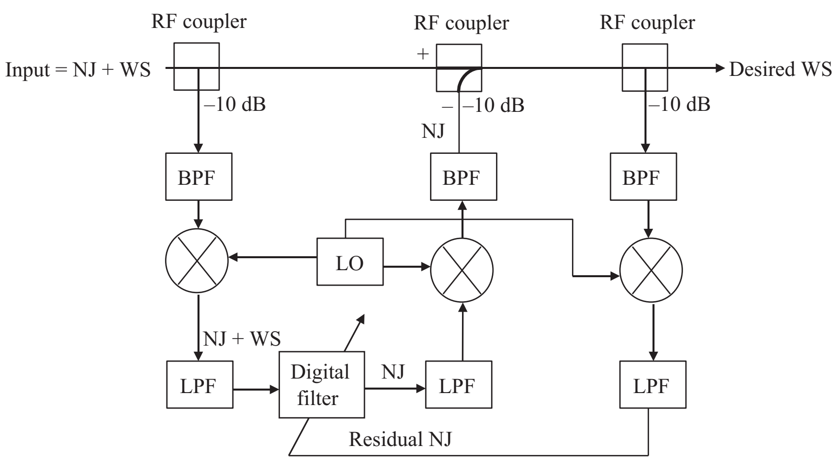


Figure 8.15 Analog/digital adaptive filter

8.2.2.5 Wideband ALE jammer suppressor adaptive filter

A wideband ALE jammer suppressor filter is shown in Figure 8.16. The adaptive filter is connected in parallel with the communication system, and the only components in-line with the system are three couplers and one amplifier. The group delay through the amplifier is constant across the band of interest, with a deviation of less than 100 ps. The amplifier is placed in the system to isolate the reference channel from the narrowband channel. This prevents the narrowband signal from feeding back into the reference channel.

The wideband high-frequency composite signal is split using a −10 dB coupler, amplified and downconverted to an IF using an adjustable synthesizer. This provides the reference signal for the digital filter. The signal is filtered, amplified, and quadrature downconverted to baseband, where the quadrature signals are filtered and amplified and provide the reference signals to the digital filters. Elliptical low-pass filters are used to achieve fast roll-offs and relatively flat group delay in the passband. The digital filters produce the estimated narrowband signals. The outputs of the digital filters are low-pass filtered, amplified, and quadrature upconverted to the IF band and summed together to eliminate the unwanted sidebands. Quadrature methods are used to cancel out the unwanted sidebands without using tunable filters for broadband applications. Figure 8.17 shows how the

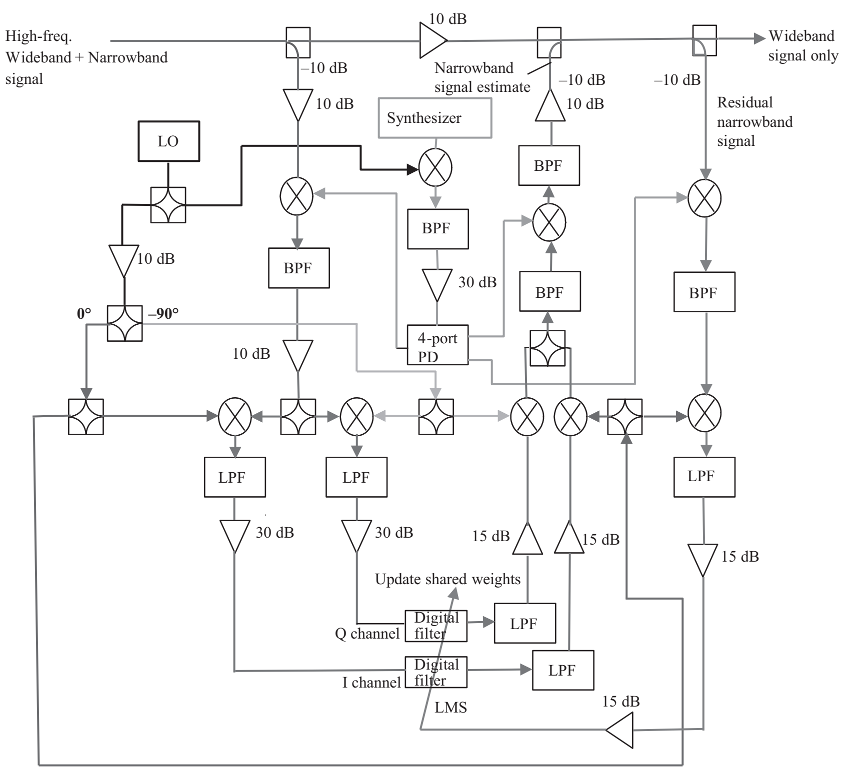


Figure 8.16 Wideband ALE jammer suppressor filter

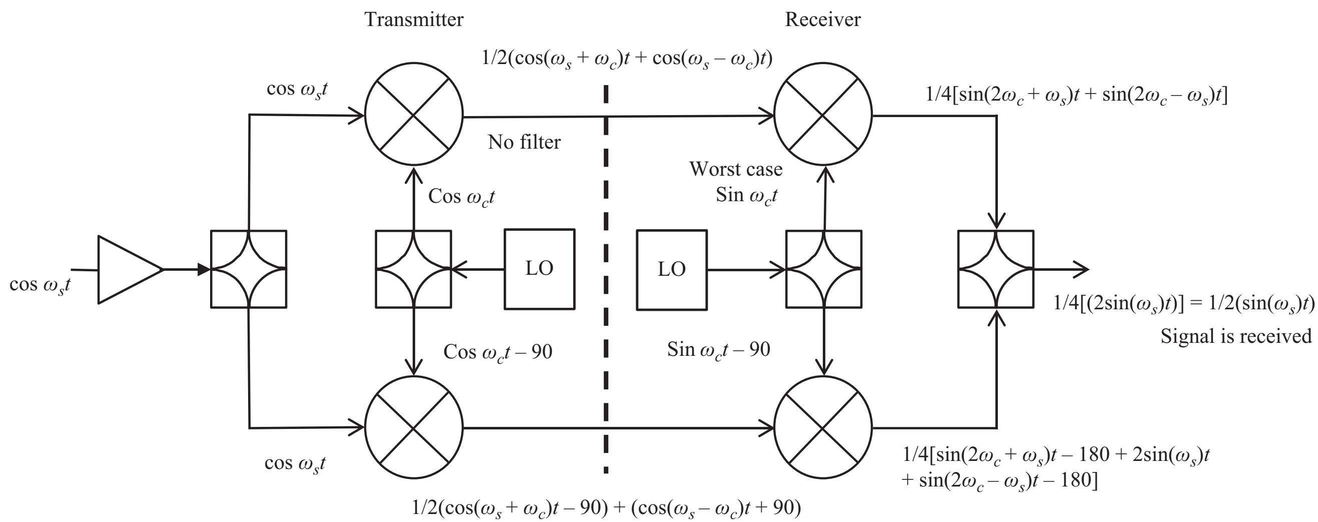


Figure 8.17 Quadrature up/down conversions eliminates S or D products without filters

unwanted sidebands during the mixer process are eliminated and the desired signal is the output. The signal is filtered and upconverted by the synthesizer to the desired frequency and then filtered, amplified, and subtracted from the composite signal to eliminate the narrowband signal. The output provides both the desired wideband signal and the error. This signal is split, and the in-phase error is downconverted in two stages to baseband to update the filter weights.

The digital filters digitize the reference input using analog-to-digital converters and feed this signal through a decorrelation delay. This decorrelates the wideband signal and allows the narrowband signal, because of its periodicity, to remain correlated. The delayed signal is fed to a tapped delay line containing 16 different time-delayed signals or taps that are multiplied by the error feedback and then accumulated to update the weight values. The analog error signal is single-bit quantized and scaled by adjusting the μ value. The μ value determines the filter's sensitivity to the error feedback. The 16 taps are then multiplied by these new weight values and converted to analog signals, where they are summed together to form the predicted narrowband output. A selectable delay was incorporated in the design to provide adjustment for the time-delay compensation. This was necessary because of the delay through the digital and analog portions of the filter. When the filter generates the predicted narrowband signal for cancelation, the error produced needs to update the portion of the signal that caused the error. If there is delay in this path, then the error will be updating a different portion of the signal from the part that generated the error. Since the analog portion produces a delay that is not quantized with regard to a certain number of clock cycles, a variable delay was designed to select small increments of a clock cycle. This allows for better resolution in selecting the correct compensation delay. A rotary switch, mounted on the board, is provided for adjusting the delay for each digital filter.

8.2.2.6 Simulation

A limited simulation of the ALE system was performed due to the complexity and large number of variables contained in the ALE system definition. Since simulation time is dependent on the sample rate of the computer clock, a baseband representation was used. The amount of computer time to do the simulation at high frequencies is impractical. Existing models were used to form the desired circuit. The frequency is swept across a 30-MHz band, and the output power is measured and displayed. Figure 8.18 shows the cancelation across the instantaneous bandwidth. The filter is unstable at the filter edges because of the phase distortion and aliasing effects. The spikes at the center frequency are a result of using only a 16-tap filter. The low-frequency components cannot be resolved in the tap delay line. The cancelation achieved in the simulation is greater than 40 dB, except around the center frequency.

8.2.2.7 Results

The adaptive filter cancelation bandwidth is similar to the results obtained in the simulation (Figure 8.19). This shows the response of the filter as a CW tone is swept across the selected band. The bandwidth is less than the simulation to

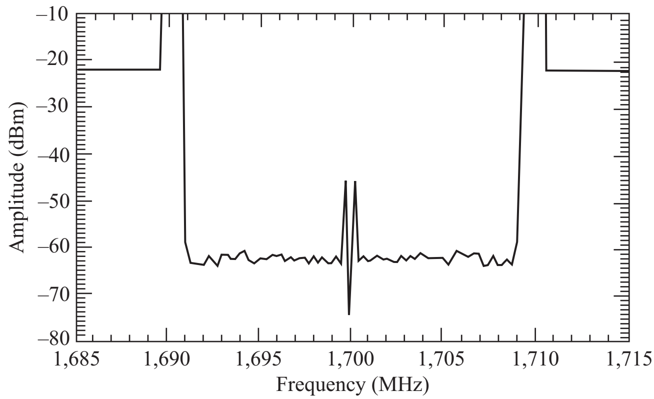


Figure 8.18 ALE simulation

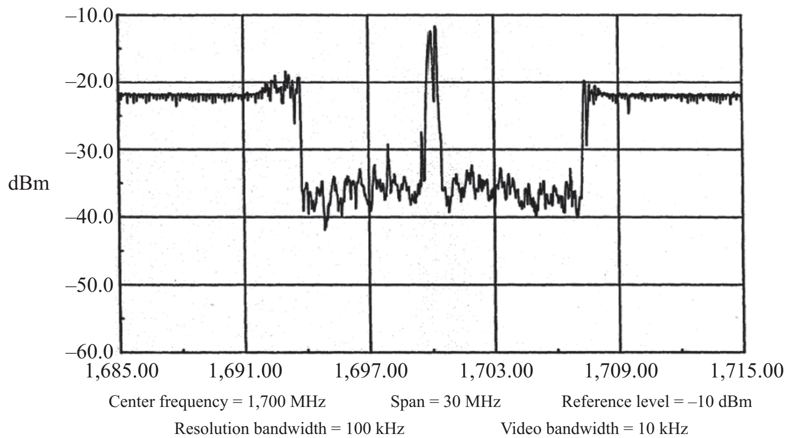


Figure 8.19 Results of filter cancelation bandwidth

suppress the phase distortion at the band edge. The center of the band is the response of the filter when it is mixed down to direct current (DC). Since the ALE is alternating current coupled in the hardware, it cannot respond to DC or low frequencies, which results in no cancelation of the signal. The amount of cancelation across the band is approximately 10 dB, with up to 30 dB at certain frequencies. This can be compared to the 40-dB cancelation across the band achieved in the simulation.

Quadrature imbalance, which is difficult to measure at high frequencies using frequency conversion processes, is a major factor in the amount of suppression of the narrowband tone. Thus, the quadrature balance was estimated and the hardware tuned to achieve maximum performance. The synthesizer can be tuned to select a

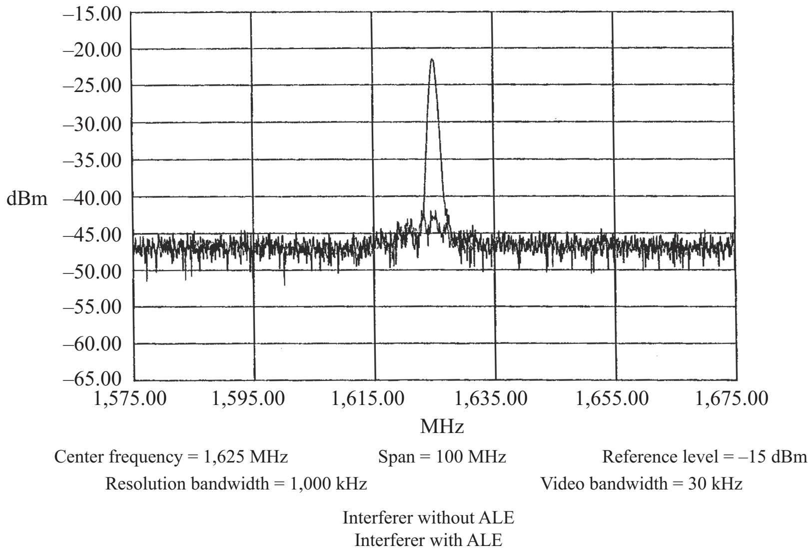


Figure 8.20 Cancelation for a single interferer

specified frequency to achieve maximum cancelation of that frequency across a given bandwidth. The amount of cancelation is dependent on many factors, such as I–Q balance for both phase and amplitude, phase linearity across the band, and stability and noise in the system.

The cancelation for a single interferer is shown in Figure 8.20. The response shows a single interferer with and without the ALE in the system. The single tone is suppressed by approximately 20 dB. ALEs can suppress more than one interferer at a time, as shown in Figure 8.21. Generally, there is degradation in performance in the amount of suppression with respect to the number of interferers. Also, more tones produce more spurious responses in the mixing processes in the system.

8.2.2.8 Amplitude and phase suppression results

The main criterion for achieving the maximum cancelation of the interferer is to ensure that its replicated waveform is exactly equal in amplitude and exactly 180° out of phase. This provides perfect cancelation. If there is an error in the amplitude or an error in the phase, this will result in a nonperfect cancelation of the waveform. The cancelation performance degrades when the amplitude is not exact (Table 8.1). In addition, if the phase of the cancelation waveform is not exactly 180°, the cancelation performance degrades (Table 8.2).

Both these sources of errors need to be considered when analyzing the amount of cancelation that can be obtained. For analog systems, these parameters are very difficult to maintain over temperature, hardware variations, and vibrations. Digital systems are much better to use for cancelation, and a higher cancelation performance can be achieved. The sooner a system can digitize the signal and perform

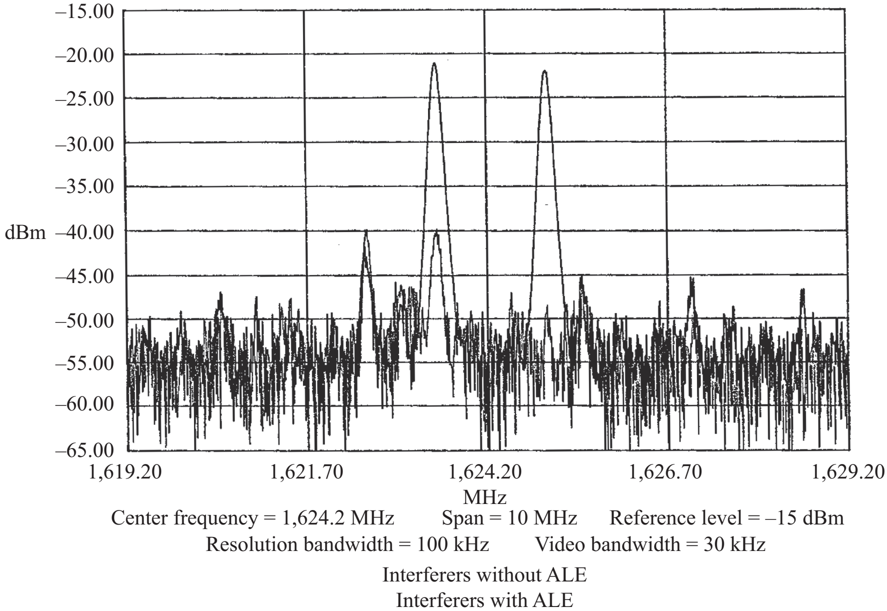


Figure 8.21 Cancelation for multiple interferers

Table 8.1 Jammer cancelation performance with amplitude variations

Suppression vs amplitude	
Amplitude ($B = 1$)	**Suppression (dB)**
1.005	52.1
1.01	46.1
1.02	40.1
1.04	34.2
1.06	30.7
1.08	28.3
1.10	26.4
1.12	24.9
1.14	23.7
1.16	22.6
1.18	21.7
1.20	20.8
1.22	20.1
1.24	19.4
1.26	18.8
1.28	18.2
1.30	17.7

Tablee 8.2 Jammer cancelation performance with phase variations

Suppression vs phase error	
Error (°)	**Suppression (dB)**
0.1	61.2
0.5	47.2
1.0	41.2
2.0	35.2
3.0	31.6
4.0	29.1
5.0	27.2
6.0	25.6
7.0	24.3
8.0	23.1
9.0	22.1
10.0	21.2
11.0	20.3
12.0	19.6
13.0	18.9
14.0	18.2
15.0	17.6

these types of functions in the digital domain, the better the performance that can generally be realized.

8.2.3 Gram–Schmidt orthogonalizer

Signals on the x- and y-axes are by definition orthogonal and in this case are 90° out of phase (Figure 8.22). All signals can be represented by the sum of weighted orthonormal functions or the magnitudes of orthonormal functions. Orthonormal functions are normalized vectors with a magnitude of one. For example, if a signal $S_1(t)$ is placed on the X_1 axis, then its value is the orthonormal function times the magnitude a. If another signal $S_1(t)$ is somewhere between the X_1 axis and the X_2 axis, then it is the sum of the magnitude b of $X_1(t)$ and the magnitude c of $X_2(t)$ (Figure 8.22):

$$S_1(t) = aX_1(t)$$
$$S_2(t) = bX_1(t) + cX_2(t)$$

where $X_1(t)$, $X_2(t)$, ... are orthonormal functions, and a, b, c, ... are the weighting coefficients.

The constant a is simply the magnitude of $S_1(t)$, since the magnitude of $X_1(t)$, by definition, is unity. To solve for b, the second equation is multiplied by $X_1(t)$ and integrated:

$$\int S_2(t)X_1(t)dt = \int bX_1(t)X_1(t)dt + \int cX_2(t)X_1(t)dt$$

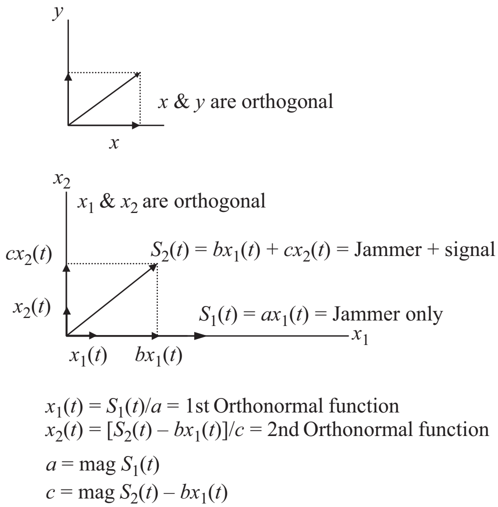


Figure 8.22 Orthonormal vectors

On the right side of this equation, the second term is equal to zero (the inner product of two orthonormal functions = 0), and the first term is equal to b (the inner product of an orthonormal function with itself = 1). Solving for b produces:

$$b = \int S_2(t)X_1(t)dt$$

This is the projection of $S_2(t)$ on $X_1(t)$.

The constant (c) is determined by the same procedure except that $X_2(t)$ is used in place of $X_1(t)$ when multiplying the second equation. Therefore,

$$\int S_2(t)X_2(t)dt = \int bX_1(t)X_2(t)dt + \int cX_2(t)X_2(t)dt$$

where $c = \int S_2(t)X_2(t)dt$. This is the projection of $S_2(t)$ on $X_2(t)$. Note that $X_2(t)$ is not generally in the direction of $S_2(t)$.

The first orthonormal basis function is therefore defined as:

$$X_1(t) = S_1(t)/a$$

where a is the magnitude of $S_1(t)$.

The second orthonormal basis function is derived by subtracting the projection of $S_2(t)$ on $X_1(t)$ from the signal $S_2(t)$ and dividing by the total magnitude:

$$X_2(t) = [S_2(t) - bX_1(t)]/c$$

where c is the magnitude of the resultant vector $S_2(t) - bX_1(t)$. Thus, the magnitudes are equal to:

$$a = S_1(t)/X_1(t) = abs(S_1)$$
$$b = \int S_2(t)X_1(t)dt \leq S_2, X_1 \geq \text{inner product}$$
$$c = \int S_2(t)X_2(t)dt \leq S_2, X_2 \geq \text{inner product}$$

A phasor diagram is provided in Figure 8.22 to show the projections of the vectors.

8.2.4 Basic GSO

A basic Gram–Schmidt orthogonalizer (GSO) system is shown in Figure 8.23. The weight (w_1) is chosen so that the two outputs, V_o and W_o, are orthogonal; that is, the inner product $\langle V_o, W_o\rangle = 0$. This gives the result:

$$\langle J + S_1, J + S_2 - w_1(J + S_1)\rangle = 0$$
$$\langle J, J\rangle(1 - w_1) - w_1\langle S_1, S_1\rangle = 0$$
$$w_1 = \frac{|J|^2}{|J|^2 + |S_1|^2} = \frac{1}{1+p}$$
$$p = \frac{1 - w_1}{w_1} = \frac{|S_1|^2}{|J|^2}$$

Note the following assumptions:

1. Same J in both inputs.
2. J, S_1, and S_2 are orthogonal.
3. $|J|^2$, $|S_1|^2$, and $|S_2|^2$ are known.
4. $p \ll 1$. This means the jammer is much larger than the signal S_1.

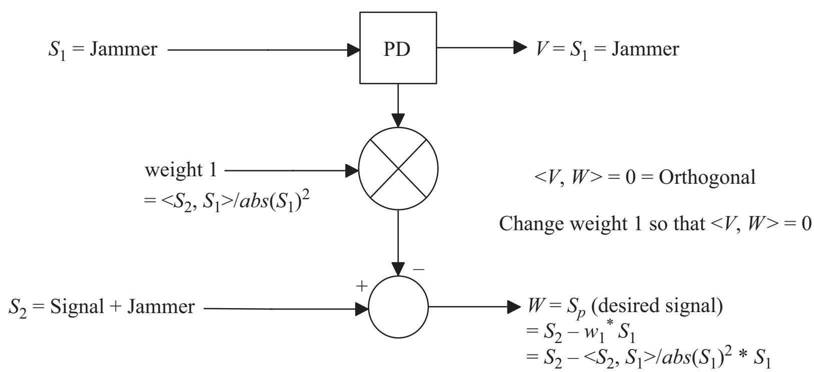


Figure 8.23 Gram–Schmidt orthogonalizer (GSO) for simple jammer suppressor

The outputs are

$$\begin{aligned} V_o &= J + S_1 \\ W_o &= J + S_2 - w_1(J + S_1) = J(1 - w_1) + S_2 - w_1 S_1 = J(p/(1+p)) \\ &\quad + S_2 - S_1(1/(1+p)) \end{aligned}$$

Since $p \ll 1$, then:

$$W_o = J_p + S_2 - S_1 = \ |S_1 - S_2|^2$$

This shows that the jammer signal has been attenuated in the W_o output.

Suppose only a jammer exists in one of the inputs, S_1, and the jammer plus signal is in S_2, as shown in Figure 8.24. Taking the projection of S_2 on the orthonormal function ($Q_1 = S_1/|S_1|$) provides the amount of jammer present in S_2:

$$b = \left\langle S_2, \frac{S_1}{|S_1|} \right\rangle$$

This scalar quantity multiplied by Q_1 produces the jammer vector bQ_1. Subtracting the jammer vector from S_2 gives the amount of signal present (S_p) in S_2:

$$S_p = S_2 - bQ_1 = S_2 - \left\langle S_2, \frac{S_1}{|S_1|} \right\rangle \frac{S_1}{|S_1|}$$

Thus, the jammer is eliminated. In the implementation of these systems, the $|S_1|$ values are combined in a scale factor k where:

$$k = 1/|S_1|^2$$

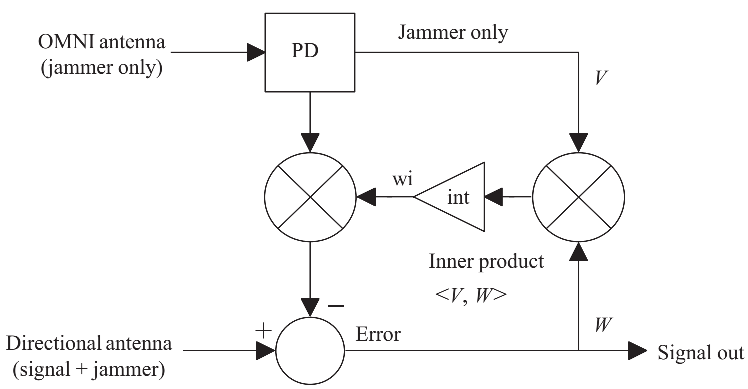


Figure 8.24 Basic GSO used for jammer suppression

The final result is then:

$$S_p = S_2 - k\langle S_2, S_1\rangle S_1$$

The constant (k) is incorporated in the specification of the weight value or in the integration process during the generation of the weights in an adaptive system.

8.2.5 Adaptive GSO implementation

If a system uses an omnidirectional antenna and assumes jammer only for S_1 (since the jammer is much larger in amplitude than the desired signal) and a directional antenna for signal plus jammer for S_2 (since the antenna will be pointed toward the desired signal), then the previous example will apply. The technique starts with updating weights to force the outputs to be orthogonal so that the inner product $\langle v, w\rangle = 0$. The weights are updated by the inner product of the outputs.

Assume only the jammer signal is $S_1(t)$. The magnitude of the jammer is equal to $a = abs\ (S_1)$. Signal $S_2(t)$ is made up of a signal vector and a jammer vector. Therefore, the magnitude of the signal vector $S_2(t)$ that contains no jammer is equal to:

$$S_v = cx_2(t) = S_2(t) - bx_1(t)$$
$$x_1(t) = S_1(t)/abs(S_1)$$
$$b = \langle S_2, x_1\rangle = \langle S_2, S_1/abs(S_1)\rangle$$
$$S_v = S_2 - [\langle S_2, S_1/abs(S_1)\rangle S_1/abs(S_1)]$$
$$S_v = S_2 - \left[\langle S_2, S_1\rangle S_1/abs(S_1)^2\right] = S_2 - S_1\left[\langle S_2, S_1\rangle/abs(S_1)^2\right]$$
$$\text{Weight value} = w = \langle S_2, S_1\rangle/abs(S_1)^2$$
$$S_v = S_2 - wS_1 = S_{\text{dir}} - w_1 S_{\text{omni}}$$

An adaptive filter configured as an adaptive noise canceller can be used as a GSO jammer suppressor (Figure 8.25).

This shows a quadrature system with separate I and Q outputs and separate weight generators. The error signal is produced by subtracting the weighted reference input signal (the received signal from the omnidirectional antenna) from the signal received from the directional antenna:

$$e = S_{\text{dir}} - w_1 S_{\text{omni}}$$

The square error is therefore:

$$e^2 = S_{\text{dir}}^2 - 2w_1 S_{\text{dir}} S_{\text{omni}} + w_1^2 S_{\text{omni}}^2$$

The MSE is:

$$\text{MSE} = e^2 = w_1^2 S_{\text{omni}}^2 - 2w_1 S_{\text{dir}} S_{\text{omni}} + S_{\text{dir}}^2$$

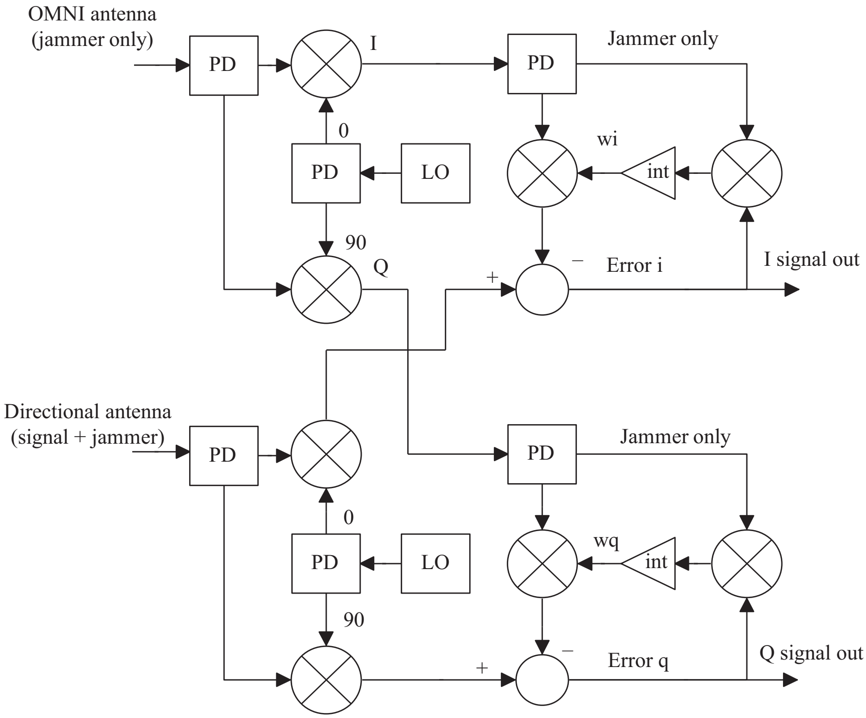


Figure 8.25 Quadrature adaptive system

Taking the gradient of the MSE and setting this equal to zero, the optimum weight value can be solved:

$$2w_1 S_{\text{omni}}^2 - 2S_{\text{dir}} S_{\text{omni}} = 0$$
$$w_{1(\text{opt})} = S_{\text{dir}} S_{\text{omni}} / S_{\text{omni}}^2$$

Since the error output is the desired signal output, then:

$$e = S_{\text{dir}} - \left(S_{\text{dir}} S_{\text{omni}} / S_{\text{omni}}^2\right) S_{\text{omni}}$$

The inner product is defined as:

$$\langle X_1, X_1 \rangle = \int X_1^2(t)dt = E\left[X_1^2(t)\right] = \overline{X_1^2(t)} = |X_1(t)|^2$$
$$\langle X_1, X_2 \rangle = \int X_1(t)X_2(t)dt = E[X_1(t)X_2(t)] = \overline{X_1(t)X_2(t)}$$

Therefore, the error can be expressed as:

$$e = S_{\text{dir}} - \left(\frac{\langle S_{\text{dir}}, S_{\text{omni}} \rangle}{|S_{\text{omni}}|^2}\right) S_{\text{omni}}$$

The second term on the right side of this equation is the projection of S_{dir} on the orthonormal function ($S_{omni}/|S_{omni}|$) times the orthonormal function. This determines the amount of jammer present in S_{dir}. This result is then subtracted from S_{dir} to achieve the desired signal (e) and eliminate the jammer. The LMS algorithm assumes that the gradient of the MSE can be estimated by the gradient of the square error, which turns out to be twice the error times the reference.

8.3 Intercept receiver comparison

Some receivers, known as intercept receivers, are designed to intercept the transmissions of an unknown transmitter. Electronic countermeasure receivers, also a type of intercept receiver, are designed to listen to broadcasts from other sources (Figure 8.26). The most common types along with their disadvantages and advantages are described below. This is not intended to be a comprehensive list but is provided to give a general idea of what type of receivers can detect the desired signal.

Crystal video detector—generally used for narrowband detection.

Advantages:

- Small in size
- Simple to design
- Low cost

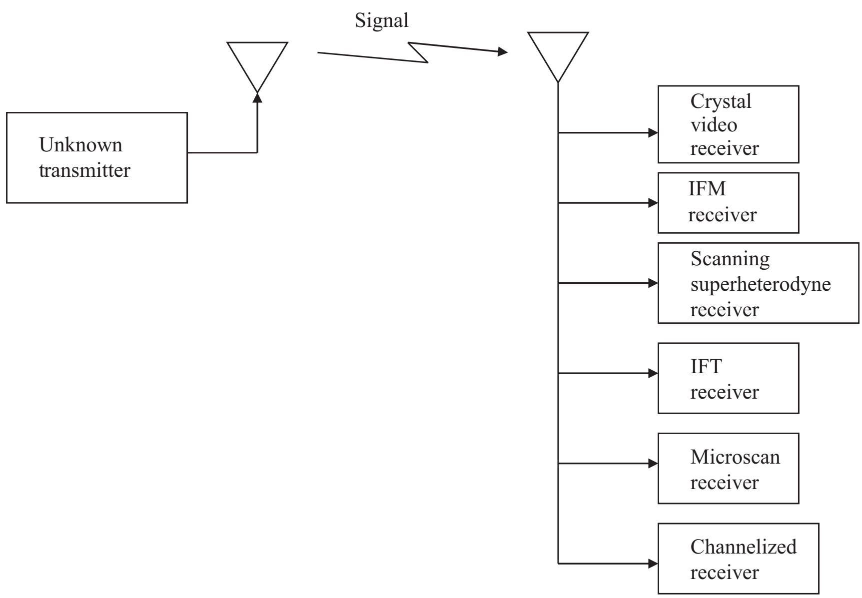


Figure 8.26 Intercept receivers for signal detection

Disadvantage:

- Narrowband detection

Instantaneous frequency measurement—delay line discriminator.
Advantages:

- Detects one-single frequency.
- High-frequency resolution detection.
- Detects wide frequency bandwidths.

Disadvantage:

- Unable to handle multiple frequencies at the same time.

Scanning superheterodyne receiver—spectrum analyzer receiver.
Advantage:

- Good frequency resolution over a wide frequency dynamic range.

Disadvantages:

- Not instantaneous
- Dependent on scan time
- Generate spurious signals

Instantaneous Fourier transform (Bragg cell)—uses acousto-optic device such as a Bragg cell where optic rays project on a surface at different points dependent on the incoming frequency.
Advantages:

- Good frequency resolution
- Multiple frequency handling
- Instantaneous wideband coverage

Disadvantages:

- Small dynamic range (20–30 dB)
- Large size of the receiver

Microscan receiver (chirped FM)—uses a surface acoustic waves device excited by an impulse input signal. Frequencies in an impulse mix with incoming signal and delayed by means of acoustic waves with the output time dependent with resultant frequency.
Advantages:

- Wideband instantaneous frequency coverage
- Good frequency sensitivity
- Very small size

Disadvantages:

- Limited dynamic range capability (30–40 dB)
- Pulse information is lost

Channelized receiver—uses multiple frequency channels providing instantaneous frequency processing.

Advantages:

- Good frequency resolution
- Detectability of multiple frequencies
- Wide frequency IDR

Disadvantage:

- Size and cost

Using intercept receivers to determine the type of jamming signal can help tremendously in deciding what type of antijam technique to use. For example, if all of the jammers are broadband, then an adaptive filter might not be the best type of antijammer to use.

8.4 Summary

Several types of jammers are introduced with advantages and disadvantages of each jammer. Burst jammers are effective at capturing the AGC time constant to maximize the ability to jam a receiver with minimum average power. The burst clamp is used to mitigate this type of jammer by preventing the AGC from being captured.

Adaptive filters can be configured as ALEs to suppress undesired narrowband signals. When these filters are used at high frequencies and over large bandwidths, modifications need to be made. The time delay through portions of the system needs to be compensated for and is accomplished in the digital filter by using a tap value to generate the signal and a delayed tap value to update the weights. Also, for small variations in delay, the clock is modified in the digital filter. Once the delay is set, constant delay over the band of operation is important for proper operation of the filter. A variable synthesizer is used in the design to achieve a wide operational bandwidth for the ALE. This allows the filter to be positioned across a given bandwidth with minimal hardware. A quadrature scheme is used to eliminate filtering constraints and provide twice the processing bandwidth for the ALE. However, the performance of the system relies on the quadrature channels being balanced in both phase and amplitude. The LO bleed-through problem is reduced by using a double down conversion scheme, since the isolation of the LO signal is greater for lower frequencies. The isolation for the conversion process is established in the first low-frequency mixer because the LO signal for the second mixer lies outside of the passband and can be filtered.

GSOs can be used to reduce the effects of jamming signals. One of the assumptions in this approach is that the jammer signal level is much higher than the desired signal level. The basic GSO has two inputs, with one containing more signal than jammer. This applies to having two antennas with one of the antennas directed toward the signal providing higher signal power. The error signal for feedback in updating the weight value is produced by subtracting the weighted

reference input signal from the received signal, which contains the higher level of desired signal. When the weight has converged, then the jamming signal is suppressed.

Spread spectrum systems can reduce a signal's detectability, and the research is on to design better intercept receivers.

8.5 Problems

1. What is meant by capturing the AGC of a system?
2. What would be a good pulse frequency for a burst jammer given an AGC response time of 1 ms?
3. What is the main difference between a digital FIR filter and an adaptive filter?
4. Why is either filtering or quadrature method required to operate the adaptive filter in the RF world?
5. What does the μ value in the LMS algorithm represent?
6. What is the result of increasing the μ value on convergence time, stability, and steady-state accuracy?
7. Why is the assumption that the jammer is the only signal present in the omnidirectional antenna of a GSO jammer suppression filter a good assumption?
8. When might the assumption that the jammer is the only signal present in the omnidirectional antenna of a GSO jammer suppression filter be a bad assumption?
9. What is the best detection receiver if cost, size, and complexity are not issues, and why?
10. If a narrowband signal is present in the desire passband, what technique can be utilized to reduce its effect on the desired signal?

Further reading

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Chapter 9
Cognitive systems

Operating communication systems in a changing environment requires the need to develop a cognitive system to mitigate the effects the environment has on communications or data links. A simple definition is as follows:

> Cognition communication is the ability for a system or systems to monitor, record, sample, test, and to be cognitive or aware of the surrounding environments; and then adapts, modifies, or changes the system to provide optimal reliable communications and communications networks. Uses reasoning and learning to further enhance performance and is proactive by predicting degradation of the link before it occurs.—Scott R. Bullock, 2014

In other words, the basic concept is to develop a system, radio, antenna, network, and then to use the available resources to monitor the environment and make an optimal change to the system to improve the wireless data link's performance.

9.1 The environment

Many elements of the environment require the need to incorporate cognitive capabilities for a communications/data link system (Figure 9.1). The following sections list the major factors, but these are not exhaustive. Many other hindrances can be present, and additional cognitive abilities can become necessary as new developments evolve and as the environment changes.

9.1.1 Jammers

One of the major environmental hindrances to any communications is jamming. This can be either unfriendly or friendly jammers. Unfriendly jammers purposely jam the data link to disable the system. They are unwanted users in the volume space. They can be very sophisticated and focused primary on jamming the signal. They can also use cognitive methods to optimize the jamming effects on the desired data link. Friendly jammers are users in the proximity of the desired data link and do not intentionally jam the desired signal. They can be other equipment and radios on the platform operating simultaneous with the data link (Figure 9.1).

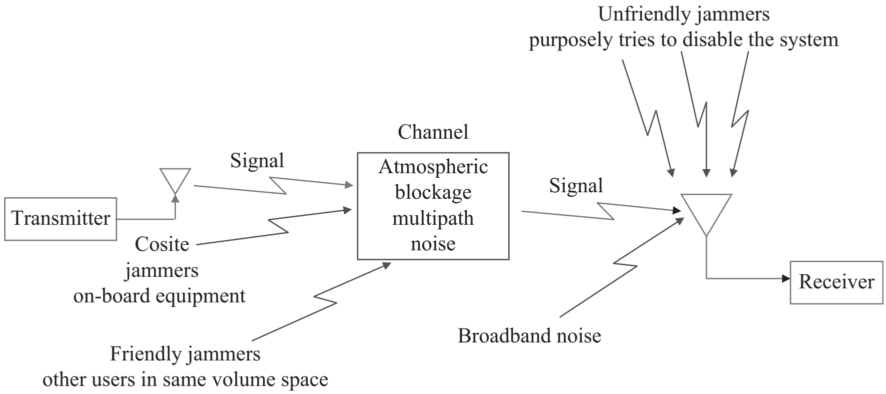


Figure 9.1 The receiver accepts both the desired signal and jammers

9.1.1.1 Unfriendly jammers

There are several types of unfriendly jammers, including barrage, pulsed, and continuous wave (CW). The type of jammer is selected to cause the greatest harm to the desired data link. Barrage jammers send out a barrage of frequencies to ensure that they jam the frequency of the desired data link. A pulsed jammer produces a very large pulse or spike that can be very detrimental to the data link and still only using minimum average power. In addition, it can capture the AGC. CW jammers can increase the power and jam the desired signal by overpowering the received signal.

For a cognitive jammer, it has the ability to assess the environment of the desired data link and select the type of jammer, the pulse duration, the amount of power required, and other parameters to improve its effectiveness. Unfriendly cognitive jammers generally pose the biggest threat to the communications link since they can adapt to the changes the receiver makes to mitigate the jammer. For example, if the receiver realizes that a pulse jammer has captured the AGC, it can change the time constant. The cognitive jammer monitors the receiver's AGC's response time and changes the pulse timing to maintain maximum jammer efficiency. So no matter what antijam process the receiver uses to mitigate the interference, the cognitive jammer adapts to the changes. In addition, a cognitive jammer can learn from the past receiver's mode of operation and then can make more intelligent changes in the future (Figure 9.1).

9.1.1.2 Friendly jammers

There are many types of friendly jammers—basically anything that is not an intentional jammer. This includes the equipment either on or in proximity to the platform that causes unintentional interference to the data link. It can also involve multiple users that are trying to communication at the same time. This equipment can be operating at the same frequency, or it can have harmonics or spurious signals that jam the data link. It also can be powerful enough to jam the data link by

saturating the front end or increasing the noise floor level, which reduces the sensitivity of the receiver. These types of jammers are often referred to as Cosite or colocated jammers, since they exist together at the site or platform of the data link. This is generally on-board equipment or radios that may have other purposes or missions (Figure 9.1).

9.1.2 Channel degradation

The channel or path of the data link can be degraded by various factors such as atmospheric changes and blockage from obstacles like hills, buildings, and multipath (Figure 9.1). All of these elements can reduce the desired signal level or increase the noise, which in turn can degrade the signal level.

In addition, broadband noise caused by adjacent equipment can degrade the data link by raising the noise floor, which causes the data link to have an insufficient signal-to-noise ratio (SNR).

9.2 Cognitive capabilities

There are three basic cognitive approaches used to analyze cognitive processes. The first one is focused only on the radio, and it is capability to change radio configurations depending on the environment. This is known as a cognitive radio (CR) and is focused on the physical or data link layers in the system protocol stack. The goals for a CR are localized only to the radios and its control functions. The second approach is expanded to use not only the physical and data link layers but uses a crosslayer design to accomplish a single goal using direct communications between nonadjacent layers. This approach uses the sharing of internal information between layers, which includes observations at one layer to provide adaptation at another layer. The third approach uses a total cognitive system or network solution which performs beyond crosslayer and considers all goals in the process. The goals of the system approach are based on end-to-end network performance using multiple layers for decision-making. It has the ability to learn from past adaptations of the entire system, which is the preferred method for applications where there are multiple mitigation capabilities.

Several cognitive techniques can mitigate the effects of jammers and channel degradation to improve the data link's performance, including dynamic spectrum access (DSA); adaptive power gain control; modification of the waveform including the order or type of modulation; spread spectrum and error correction; adaptive filters; Cosite radio frequency (RF) tunable filters; dynamic antenna techniques using active electronically scanned arrays (AESAs) such as multiple-in and multiple-out (MIMO) antenna systems; and network reconfigurations such as multihop, ad hoc meshed networks, self-forming, and self-healing.

A software-defined radio (SDR) has the ability to change the capabilities of the radio by simply changing the software during operation and is a key component in many of the cognitive solutions (Figure 9.2).

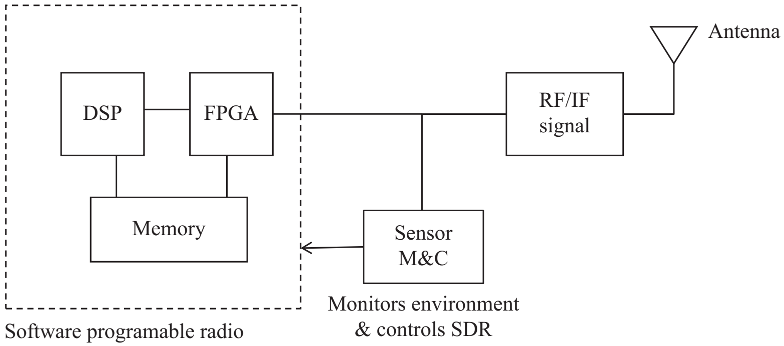


Figure 9.2 Cognitive hardware and software-defined radios (SDRs)

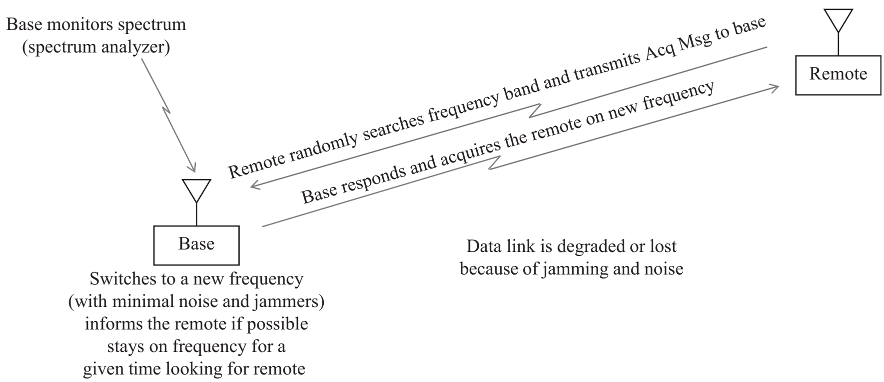


Figure 9.3 Dynamic spectrum access (DSA)

9.2.1 Dynamic spectrum access

The cognitive capability of DSA is focused on the frequency of operation of the data link. This is generally included in the CR, but the software resource manager controller can also handle it using a frequency synthesizer in the front end before the radio. The system is capable of scanning the frequency spectrum (miniature spectrum analyzer) to be cognitive of the frequencies and use of the spectrum to make an intelligent decision of what frequency is optimal for use by the data link. The cognitive system continuously monitors or pulses the environment between transmissions to evaluate the spectrum and uses that information to determine the best frequencies with minimal noise and jamming for the data link (Figure 9.3).

This requires system and network coordination to switch frequencies between two users or multiple nodes in a network.

A simple example of DSA between two users is as follows:

1. The data link exhibits poor performance or link loss.
2. The base informs the remote of the desired frequency if possible.

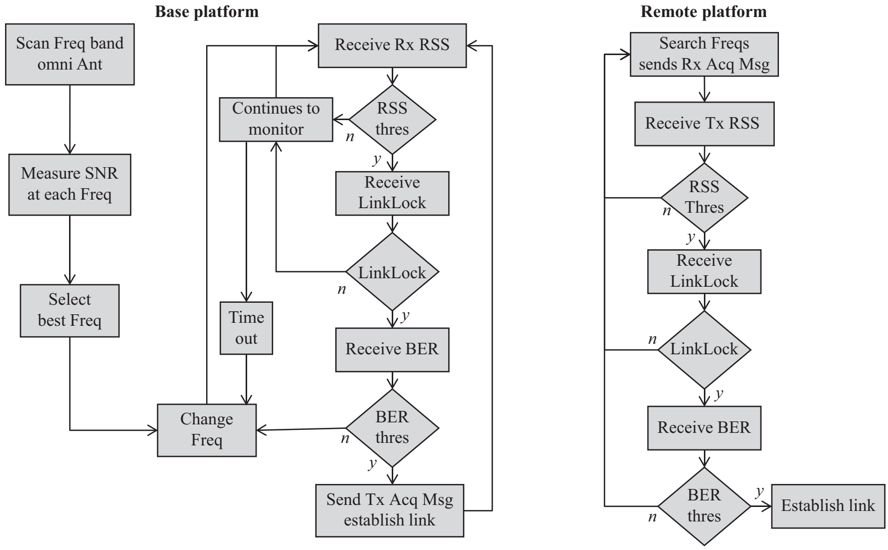


Figure 9.4 Dynamic frequency allocation between the base and the remote

3. The base switches to an unused frequency with minimal noise and jamming, or "white space."
4. The base stays on that frequency for a specified time looking for the remote's signal.
5. The remote searches for the base's new frequency using a random search of possible frequencies.
6. The remote sends a short acquisition sequence on each frequency and listens for a response.
7. The base responds with a short acquisition sequence upon detection of the remote's signal.
8. The process repeats until acquisition is successful.

The example is shown in Figure 9.3. In addition, a flow diagram shows the dynamic spectrum allocation process between a base and a remote (Figure 9.4).

9.2.1.1 Alternative to implementing DSA-using random frequency switching

Another alternative to DSA is where the base and the remote units have a known random switching frequency sequence. This is known and stored into memory of both systems and can be changed or reprogramed into the devices before the mission. During communications between the systems, this random frequency sequence can be changed or adapted to the environment.

The process of events of the alternative DSA is as follows:

1. The data link exhibits poor performance or link loss.
2. The base switches to the next frequency in the known pseudorandom sequence and stays there for a specified period of time.

3. The base informs the remote of the next frequency if possible.
4. The remote switches to the next frequency in the known random sequence and sends short acknowledgment sequence back to the base.
5. The base responds with a short acknowledgment sequence upon detection of the remote's signal and acquires the remote.
6. The process is repeated until acquisition is successful.
7. Over a period of time or number of unsuccessful tries, the system reverts to the previous DSA process.

This method is shown in Figure 9.5. The disadvantage of this process is that it does not use the cognitive capability and actually can switch to a worse frequency until it finds a good frequency. In addition, there is a probability that the systems will become out of sync and will have to revert to the first DSA random search process. However, the advantage of this DSA process is that the search time for the remote is significantly reduced.

9.2.2 Adaptive power/gain control

There are three basic methods for implementing adaptive power/gain control, which are as follows:

1. Adapt the power/gain based on the received signal strength (RSS) and bit error rate (BER)/LinkLock.
2. Adapt the power/gain level based on the range using the navigation data either from an external source or internal to the data link.
3. Use an integrated solution that combines both of these approaches using the range for coarse control and the RSS/BER/LinkLock for the fine control.

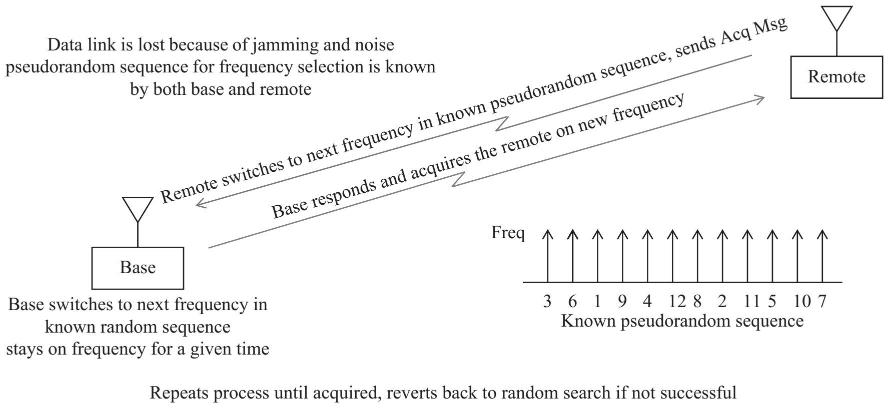


Figure 9.5 Dynamic spectrum allocation alternative

9.2.2.1 Standard closed-loop power/gain control requires modifications for convergence with two independent users

Standard power/gain control for two users will not converge to a valid result. This is because each user is setting a threshold separately (Figure 9.6). For example, the base receives a power-level RSS which is above the threshold so it reduces the power output to the remote. The remote receives a lower RSS below the threshold from the base and increases its power. The base receives a higher RSS above the threshold from the remote and decreases the power. The remote receives a lower RSS below the threshold from the base and increases the power. Therefore, the standard power control system does not converge, and a modified version of the closed-loop power control needs to be implemented.

Modified adaptive power/gain control using RSS and BER/LinkLock

This method uses the measured RSS to determine the amount of gain control in a closed-loop system with the base setting the threshold and the remote uses the change in RSS with reverse logic. Often times, the BER or LinkLock indicator is used to verify that the RSS is the desired remote user and not a jammer or other unwanted signal measurement.

The base station monitors the RSS from the remote, and if it is lower than a threshold, the base increases the power/gain of its output signal to the remote.

The remote signal measures the RSS and compares it with the previous RSS from the base to see if there is an increase or decrease in signal level. If the signal level increases, then the remote station increases its signal level by raising it power/gain of its output signal to the base station. This is somewhat counterintuitive to the standard adaptive power control. The remote looks at the delta RSS received from the base for power/gain control (Figure 9.7).

If the base station receives an RSS greater than the threshold it has set, then it lowers its power/gain to decrease the signal sent to the remote. Upon reception of

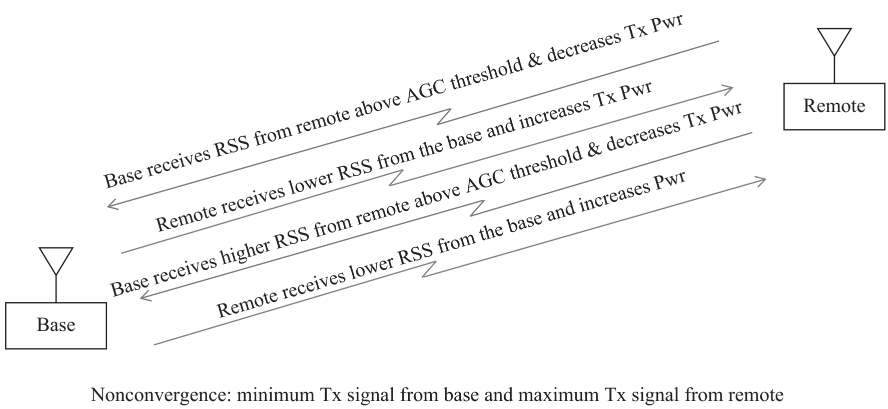


Figure 9.6 Standard closed-loop adaptive gain/power with typical AGC response of base and remote

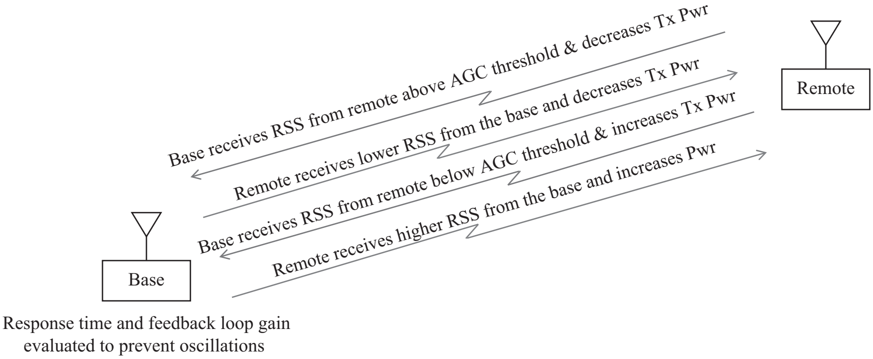


Figure 9.7 Adaptive power control between the base user and the remote user

the lower RSS from the base station, the remote station lowers its power/gain to send a lower power signal to the base station (Figure 9.7).

The base station sets the power control for each of the remote users using a threshold. This can be a fixed threshold for each of the users or a dynamic threshold for each of the users if the environments including noise and jammer levels are constantly changing. A summary of the tasks for both the base and the remote stations are as follows:

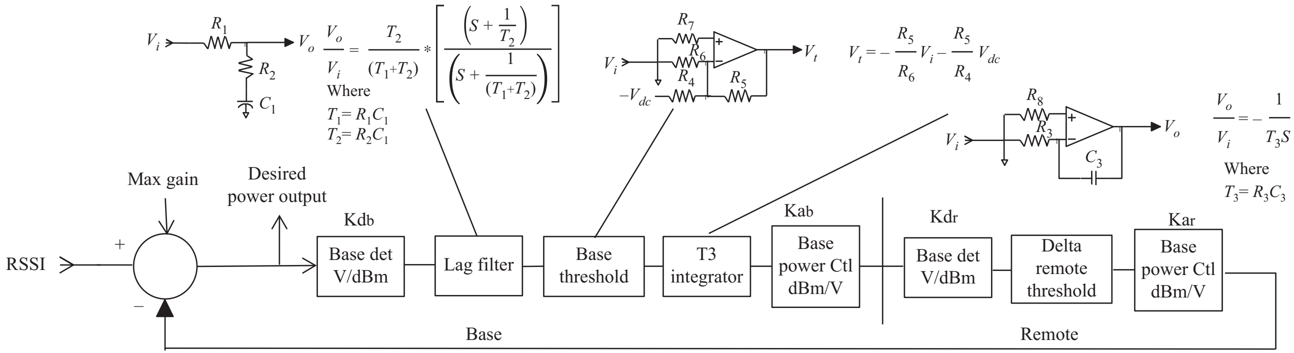
Base platform:
Receives RSS from the remote
Receives BER/LinkLock
High RSS: reduces power output
Low RSS: increases power output
Sets AGC threshold level
Standard power control/gain adjust

Remote platform:
Receives RSS from the base
Receives BER/LinkLock
Delta lower RSS: reduces power output
Delta higher RSS: increases power output
Adjusts on delta RSS level

Control theory design for a stable system

Using this method requires basic control theory since it is a feedback closed-loop system and needs to be designed to prevent oscillations. The loop incorporates both the base and the remote stations.

The closed-loop feedback system is shown in Figure 9.8. The system uses time constants for each of the blocks and includes both a loop filter to establish a pole and a zero for the root locus to ensure stability and also an integrator for producing a zero steady-state error for a step response. See Chapter 4 for detailed analysis of control systems for AGCs. The example includes the integrator, loop filter, and a fixed threshold (can be variable for more agile system). In a practical case, multiple RSSs are averaged for a given output value and are verified with a valid LinkLock



Base and remote second-order AGC closed-loop response

$$\text{Open-loop transfer function} = T_{sol} = G(s) = Kdb * Kab * Kdr * Kar * \frac{T_2}{(T_1+T_2)} \frac{\left(S+\frac{1}{T_2}\right)}{\left(S+\frac{1}{T_1+T_2}\right)} * \frac{1}{T_3S} = \frac{K\left(S+\frac{1}{T_2}\right)}{S\left(S+\frac{1}{T_1+T_2}\right)}$$

Where $K = \text{Kdb*Kab*Kdr*Kar*}(T_2/[T_1+T_2])*(1/T_3)$

$$\text{Closed-loop transfer function} = T_{scl} = G(s) / [1 + G(s)H(s)] = \frac{K\left(S+\frac{1}{T_2}\right)}{S^2+\left(\frac{1}{T_1+T_2}+K\right)S+\frac{K}{T_2}}$$

Figure 9.8 Closed-loop analysis using control theory to ensure stability

indicator. The overall response time required needs to be slower than the dynamics of the averaged RSS.

Description of the closed-loop analysis for a base and remote power control

As the RSS is received at the base, it is detected to produce a voltage level responding to the power of the RSS. This voltage is fed through a lag filter to generate poles and zeros to stabilize the control loop. The base threshold sets the voltage-level threshold, responding to the desired received signal level that is optimal for the detection of the signal. The integrator is used to ensure that there is zero steady-state error and that it holds the desired receive signal exactly at the correct level. If there is an error from the threshold, at steady state, it will be amplified by theoretical infinity to ensure no errors when the input level is not changing. The voltage-controlled amplifier converts the voltage level to a gain level, which is subtracted from the maximum gain to a value that amplifies the RSS signal to the desired signal strength set by the threshold.

With the addition of the remote in the feedback loop, the gain of the detector and the gain of the power control are added to the forward path of the feedback loop. The delta threshold adjusts the level depending on the previous level.

For the base, a large signal input causes a large signal gain, which is subtracted from the maximum gain, so a large signal reduces the gain. Adding the remote function, this large gain is further amplified by the gains in the remote, which further reduces the gain from the maximum gain.

So in summary, a large signal into the base reduces the signal-level output, and this delta reduced signal into the remote reduces the signal further, which is an extra gain toward a smaller signal or gain.

The loop shows that the larger the RSS, the larger the gain for both the base and remote and therefore the smaller the overall gain since it is subtracted from maximum gain in the base. If it was just the base, then the reduction and feedback loop would end at the dotted line (Figure 9.8). With the remote as part of the feedback loop, it increases the gain which reduces the overall gain response much faster than just the base.

In the actual performance, when the signal is high at the input to the base, the output of the base reduces the signal level and the remote detects this reduction of signal level and further lowers the signal level, which is accomplished in the total feedback loop (Figure 9.8).

Control theory feedback analysis determines the loop filter parameters to prevent oscillations. ζ is chosen for a 5% overshoot, which is a trade-off between response time and settling time of the control. The overall response time is chosen at 10 rad/s, which provides ample time for power control. This allows the response time to be fast enough for high aircraft maneuvers and also provides the time to complete the power control in the total loop response. The root locus plot ensures the stability of the system. A Bode plot analysis is also included to verify a stable closed-loop system. The Bode plot determines the gain and phase margin of the system as an indicator of stability. For further information regarding control theory and Bode plots, please refer to Chapter 4.

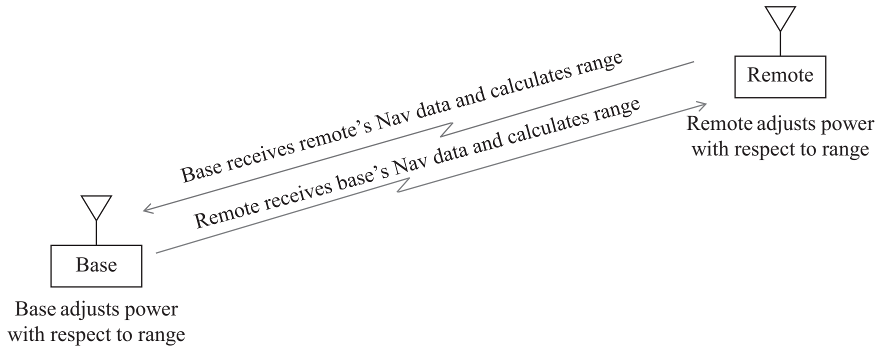


Figure 9.9 Adaptive power control between a base and a remote using Nav data for range

9.2.2.2 Adaptive power-using navigation data (external or internal) for range

The range can be calculated using the available navigation data, such as an input from an external source like a global positioning system (GPS), or it can be included in the data link itself, such as the internal data message of the common data link (CDL). The process is shown in Figure 9.9.

For the range solution, a simple stepped power control can be used. An example of a simple stepped range solution using range thresholds and power steps is shown below:

Simple stepped power control for range data:
If range is greater than 20 nmi, set to full power.
If range is less than 20 nmi, reduce power by 10 dB.
If range is less than 5 nmi, reduce power by another 10 dB.
Add hysteresis (1–5 nmi) to prevent continual switching at the transition.
Note: Finer step sizes can be used depending on the dynamics of the system.

A summary of the tasks for both the base and the remote stations are as follows:

Base platform:
Calculates range from navigation data
Adjusts power with respect to range
Sends the power level to the remote
If range is not available, use closed-loop analysis

Remote platform:
Calculates range from the navigation data
Adjusts power with respect to range
Sends the power level to the base
If range is not available, use closed-loop analysis

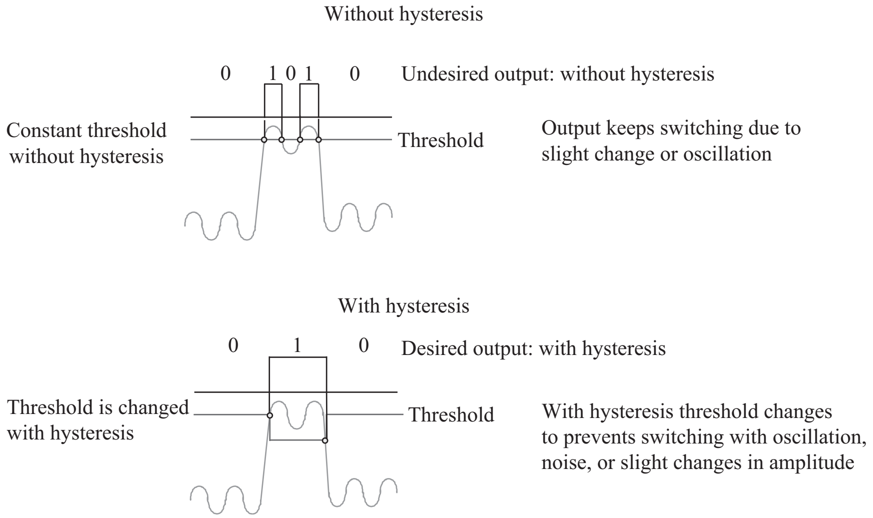


Figure 9.10 Hysteresis prevents false switching

Hysteresis prevents false switching

Using a fixed threshold in the presence of noise, oscillations, or other variations causes constant switching as the signal is affected by these distortions. In order to prevent constant switching around the threshold, hysteresis is added to the threshold. With hysteresis, the threshold level changes to prevent switching with oscillation, noise, or slight changes in amplitude (Figure 9.10).

9.2.2.3 Integrated solution of closed loop and range

This solution combines the closed-loop RSS and BER/LinkLock with the range solution using navigation data. The navigation data for range is used for a coarse power control (e.g., 10 dB steps), while the closed-loop RSS and BER/LinkLock method is used for fine power control (continuous adjustments) (Figure 9.11).

A summary of the tasks for both the base and the remote stations are as follows:

Base platform:
Calculates range from NAV data
Adjusts coarse power output with respect to range
Receives RSS from the remote
Receives BER/LinkLock
High RSS: reduces power output
Low RSS: increases power output
Fine power output adjustment
Sets AGC threshold level
Standard power control/gain adjust

Aircraft platform:
Calculates range from NAV data
Adjusts coarse power output with range
Receives RSS from the base
Receives BER/LinkLock
Delta Lower RSS: reduces power output
Delta Higher RSS: increases power output
Fine power output adjustment
Adjusts on delta RSS level

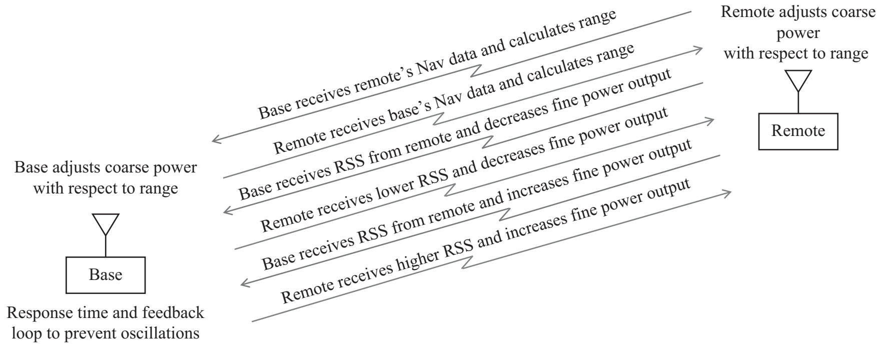


Figure 9.11 Integrated solution for adaptive power control

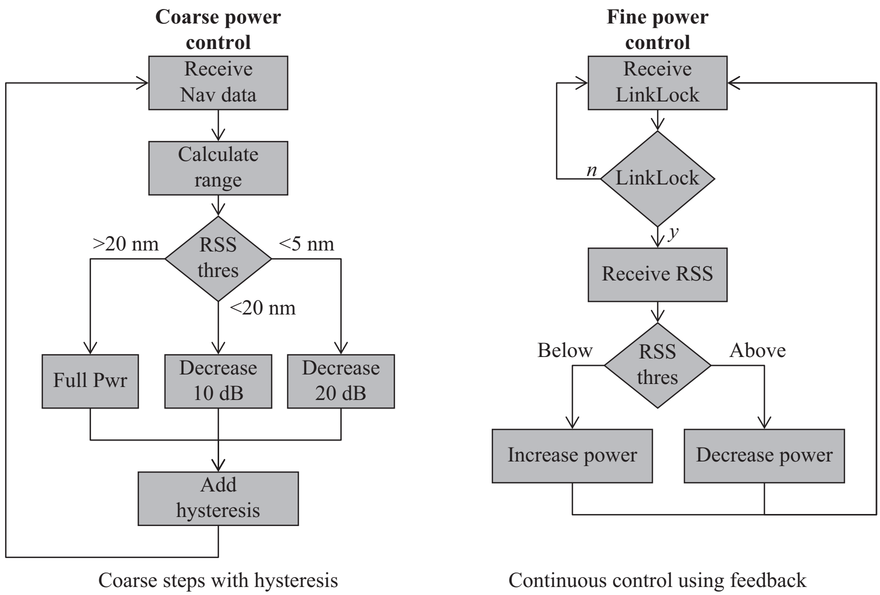


Figure 9.12 Base station adaptive power control flow diagram

Flow diagrams show the adaptive power control process between a base station and a remote station (Figures 9.12 and 9.13).

Another alternative in cellular communications is to allow the base to send out power control information to multiple users, telling them what power they should use at the measured range. This assumes that the link is available and that it has the means of getting this message to the remote stations. This same technique could be used in this application but is not as highly dynamic as the closed-loop process.

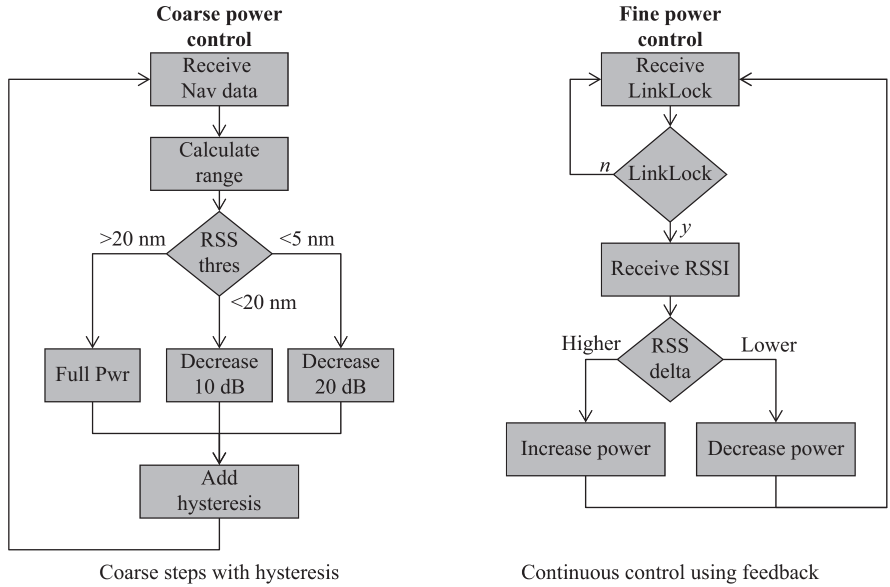


Figure 9.13 Remote station adaptive power control flow diagram

9.2.3 Cognitive techniques using modulation waveforms

Trade-offs between modulation types can be adapted to the changing environments using SDR techniques. The ability to change modulations in real time as the environment changes provides a unique cognitive solution. The modulation can be changed based on the SNR of the received signal. In a clean environment, where the SNR is high, a higher order modulation can be used to increase the data throughput. If the environment changes to a noisy jamming environment, which reduces the SNR or signal-to-jammer ratio (SJR), then a more robust, lower order modulation type can be used. The noise or jammer immunity for two different types of modulations is shown in Figure 9.14.

In a clean environment, the modulation with the maximum data rate is used or 16-state quadrature amplitude modulation (16-QAM). In noisy and jamming environments, the system adapts by switching to lower order modulation with better noise and jammer immunity at the expense of lower data rates, or binary phase-shift keying (BPSK). This is a trade-off between data rate and antijam capability: with higher data rate, there is less antijam; with lower data rate, there is more antijam. The threshold for the SNR to switch the system to a different modulation is either set or can be an adaptive threshold, which has higher performance at the cost of complexity.

The 16-QAM modulation provides four times the data rate compared with BPSK. However, it is affected more from a noisy environment and is much harder to detect without errors. The BPSK modulation is much more robust to a noisy environment and is easily detected but has a lower data rate (Figure 9.14). So the

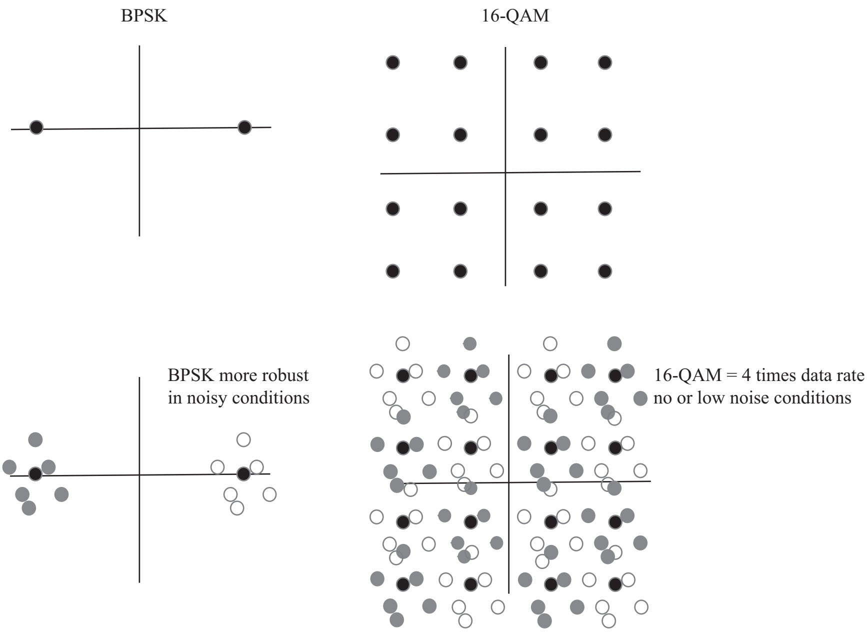


Figure 9.14 Noise immunity between BPSK and 16-QAM

cognitive solution adjusts the modulation according to the conditions of the environment. With an SDR, both modulations are available, and the type of modulation is selected very quickly. Other modulations can be made available for selection as needed. The required E_b/N_o for several types of modulations is shown in Figure 9.15.

9.2.4 Spread spectrum SS for increased process gain against jammers

Spread spectrum is used to mitigate jammers and increases the bandwidth to provide process gain. Increases the SS provides better antijam at the expense of spreading losses to the data link. If the data link is operating in a harsh environment with jammers, more spread spectrum can be applied. If the data link is operating in an environment that is free from jammers and interference, then minimal or no spread spectrum is required which eliminates the spreading loss associated with SS. The main two types of spread spectrum are direct sequence and frequency hop. Either of these can be changed rapidly with an SDR for use in a cognitive system. The system determines the amount of jamming by monitoring the spectrum, using the SJR, or BER. An example of spread spectrum is shown in Figure 9.16. The narrowband-desired signal is spread via a fast pseudonoise (PN) code and sent out of the transmitter as a spread spectrum signal. Both the spread spectrum signal and the narrowband jammer are then delivered to the receiver. The receiver contains the same fast PN code as the transmitter, and when the codes are aligned, the design

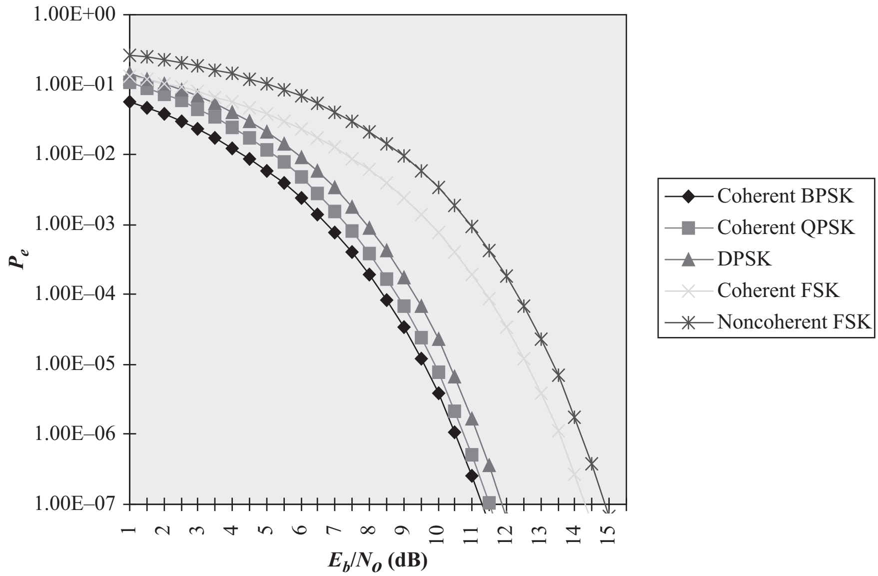


Figure 9.15 Required E_b/N_o for a given probability of error for different modulations

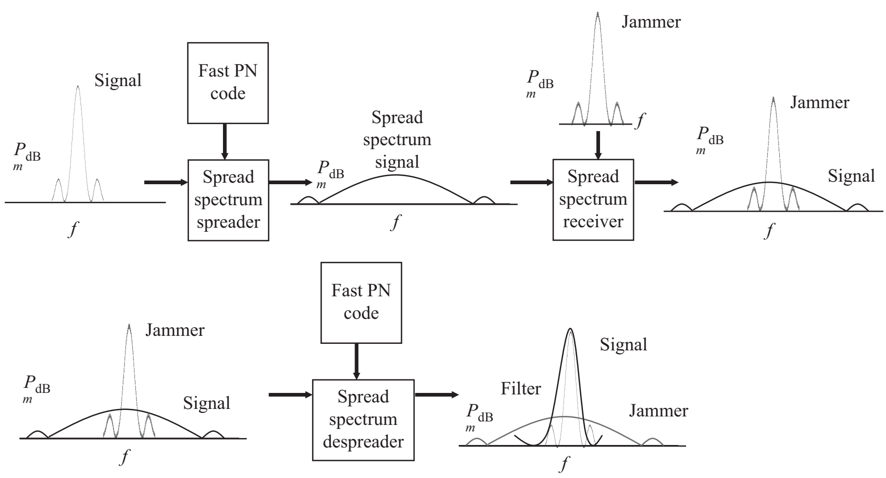


Figure 9.16 Spread spectrum techniques used only in high-jamming environments

signal is despread to the narrowband-desired signal. At the same time, the fast PN code in the receiver spreads out the jammer so that the signal is now much higher than the jammer and can be easily filtered and detected (Figure 9.16). In addition, the spread spectrum code can be adaptive or changed by simply changing a variable delay in a code generation (Figure 9.17).

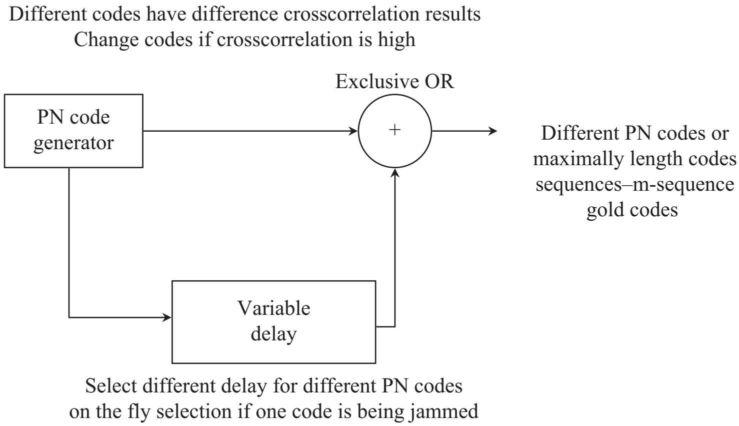


Figure 9.17 Adaptive selection of different PN codes reduces crosscorrelation jammers

9.2.5 Adaptive error correction

Error correction can also be altered depending on the jamming environment. The more error correction, the better the antijam performance at the expense of a decreased data throughput, since it requires additional bits to provide error correction. If the environment is clear, with low jamming levels, then the system can operate with less to no error correction, which increases the data throughput. If the environment is noisy with jammers, it requires more error correction at the cost of reducing the data rate. This also can be cognitive with the determination of the jamming environment similar to the spread spectrum scenario.

9.2.6 Adaptive filter for jammer mitigation

The adaptive filter can reduce narrowband jammers and changes its response as the jammer changes frequencies. This is discussed in detail in Chapter 8. By its nature, the filter is adaptive in frequency. If the environment contains a high level of narrowband interference, the adaptive filter can be incorporated into the receiver to eliminate these narrowband jammers. If the narrowband interference is not present, the adaptive filter can be disabled to improve the processing time of the received signal. In addition, the number of weights or the length of the tapped delay line can also be adaptive dependent on the type of interference. If the jammers are CW, then the number of weights and length of the tapped delay line can be reduced to improve process time. If the jammers are broader band, then more weights and longer tapped delay lines are required.

A digital adaptive filter automatically adjusts to a changing narrowband jammer. It uses an adaptive line enhancer to obtain the narrowband jammer, which in turn is used to cancel the unwanted jammer (Figure 9.18).

The decorrelation delay is the key to the operation of the adaptive filter since it separates the narrowband jammer from the wideband signal. The adaptive filter has

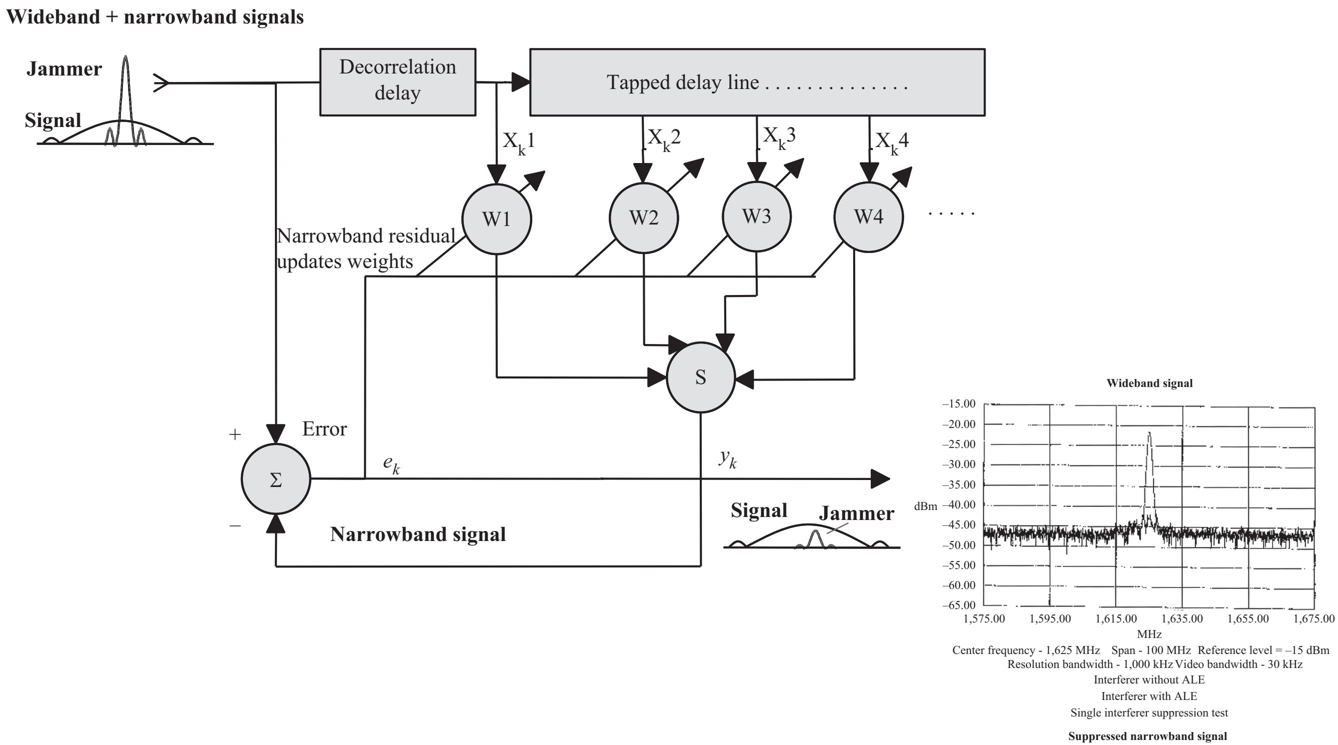


Figure 9.18 Use digital adaptive filter with a wideband signal when narrowband jammers are present

the ability to cancel jammers in both the digital domain and also at RF or intermediate frequency (IF) if needed with degraded performance.

Other adaptive methods for reducing unwanted jamming signals include the Graham–Schmidt orthogonalizer (GSO; see Chapter 8) and Cosite RF filtering.

9.2.7 Dynamic antenna techniques using AESAs

Several techniques can be included in the cognitive system with AESAs. Many of these can be employed with mechanically steered antennas, but the AESA provides a much improved means to accomplish these dynamic capabilities.

9.2.7.1 Beam steering

Beam forming and steering can be accomplished quickly using an AESA. The technique is to adapt the pointing angle of the AESA to steer away from the jammer and point it in the direction of the desired user (Figure 9.19). This is dependent on the beamwidth and how far spatially the jammer is from the desired user. The system determines the position of the jammer and the position of the user and then points the beam in the direction with the best signal from the desired user and the minimum signal from the jammer.

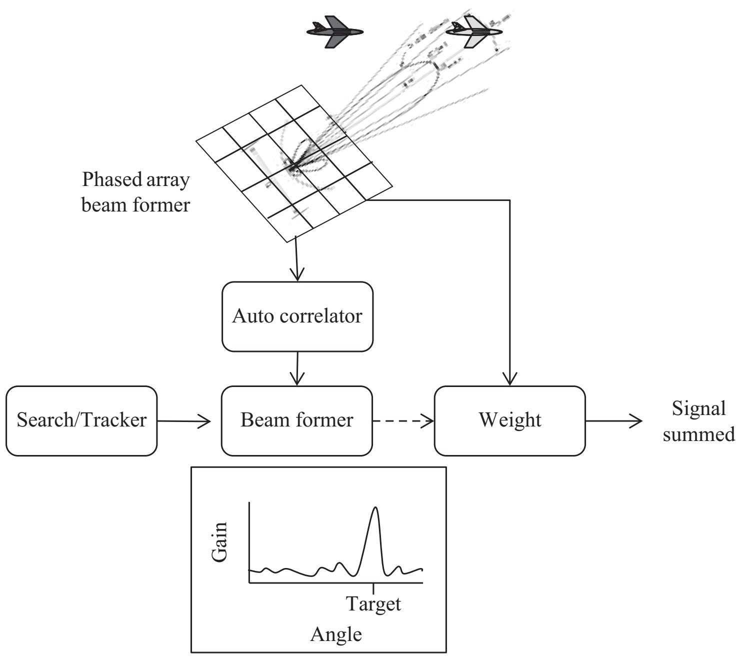


Figure 9.19 Beam former steers the AESA beam toward the desired user and away from the jammer

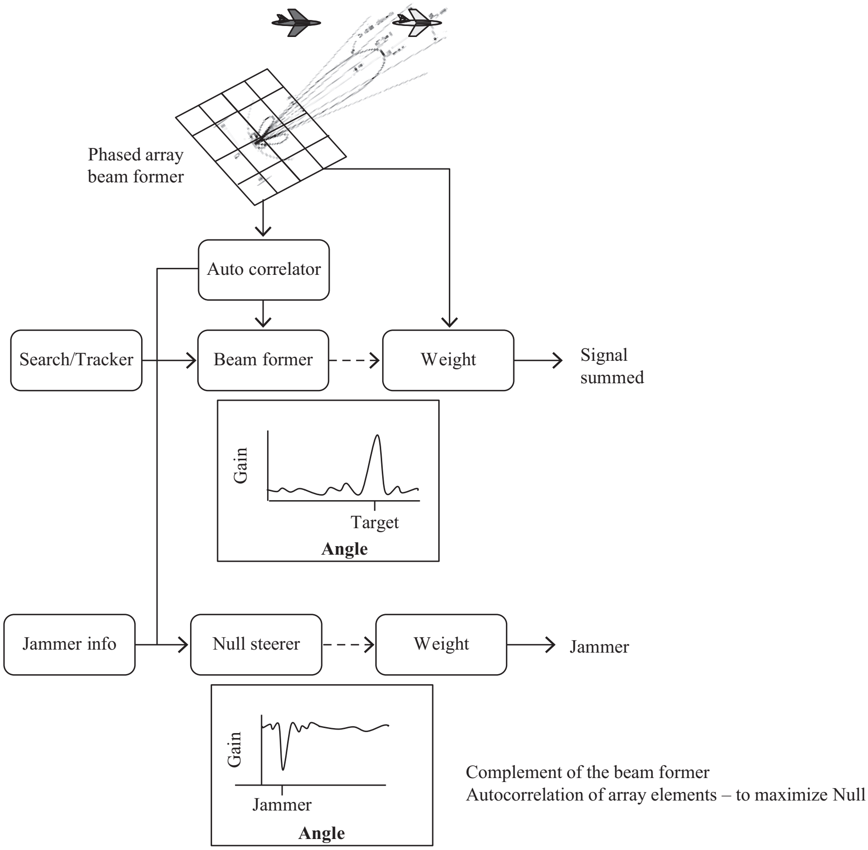


Figure 9.20 Null steering creates and steers a null of the AESA toward the jammer

9.2.7.2 Null steering

AESAs are versatile and can create a null at the location of the jammer. The system determines where the jammer is spatially located, and the AESA steers a null in the direction of the jammer to reduce the jammer level into the receiver (Figure 9.20).

9.2.7.3 Beam spoiling and narrowing

Beam spoiling involves widening the beamwidth. Beam spoiling and beam narrowing can be easily accomplished using an AESA (Figure 9.21). The narrow beam is better for power and improved spatial performance over jammers. The wide beam is better for wider coverage and better volume-search performance. The cognitive AESA adapts to the desired beamwidth for coverage versus jamming performance. Beam spoiling reduces search time but lowers the link margin and increases vulnerability to jammers. The system reviews the jamming environment and system requirements then determines whether a narrow beam is required for

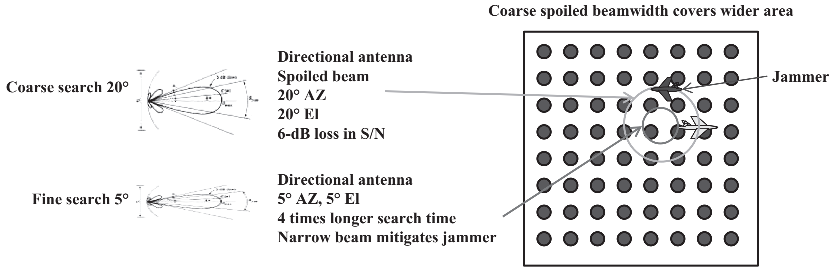
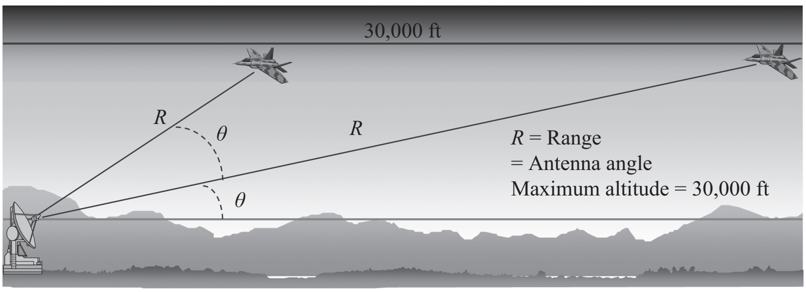


Figure 9.21 Adaptive beam spoiling and beam narrowing provides coverage and antijam trade-offs



Adapt power according to antenna angle

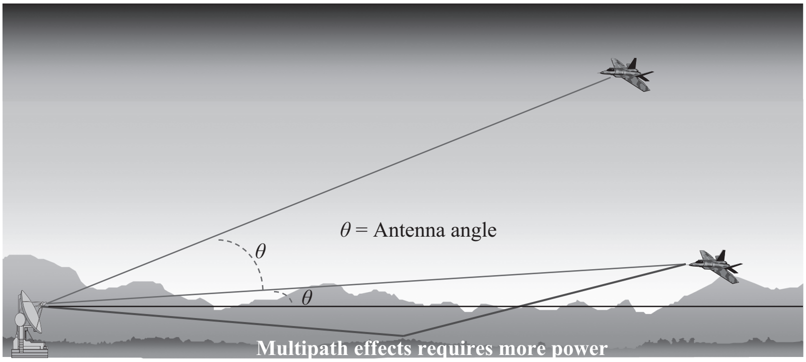
Figure 9.22 Angle of antenna determines maximum range

interference rejection or a wider beam can be used for improved coverage performance.

9.2.8 Antenna angle

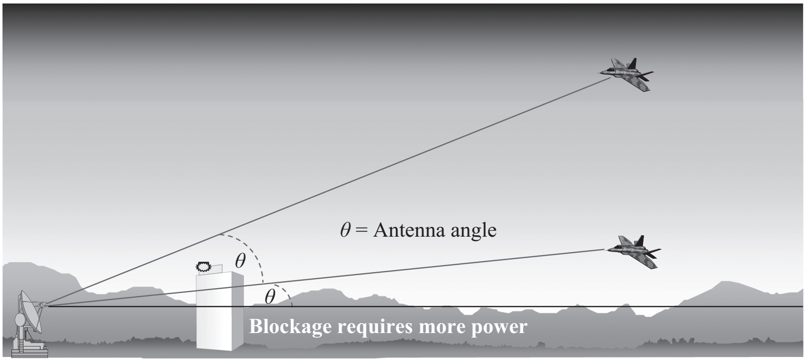
Angle of the antenna provides information to determine the maximum range, since the maximum altitude is limited by the type of aircraft (Figure 9.22). Therefore, if the elevation angle of the antenna is large, then the maximum range will be reduced which provides an opportunity to reduce the power needed for a particular target. In addition, for low elevation angles the probability of multipath effects and blockage effects are high, whereas at high elevation antenna angles, the probability of multipath and blockage effects are reduced which also is a parameter to reduce the needed power for reliable communications (Figures 9.23 and 9.24).

Multipath is generally the unwanted path that interferes with the desired direct path of the signal. However, it can be used as the desired path if the direct path is either blocked or jammed. It may be the only feasible path to the target (Figure 9.25). This technique can be useful to transmit around obstructions or to point the antenna away from the source of jamming. It uses the alternate multipath of the signal for the communications path. The system can determine if there is blockage



Adapt power according to antenna angle with multipath effects

Figure 9.23 Multipath effects increase with lower angle of antenna



Adapt power according to antenna angle with blockage effects

Figure 9.24 Blockage increase with lower angle of antenna

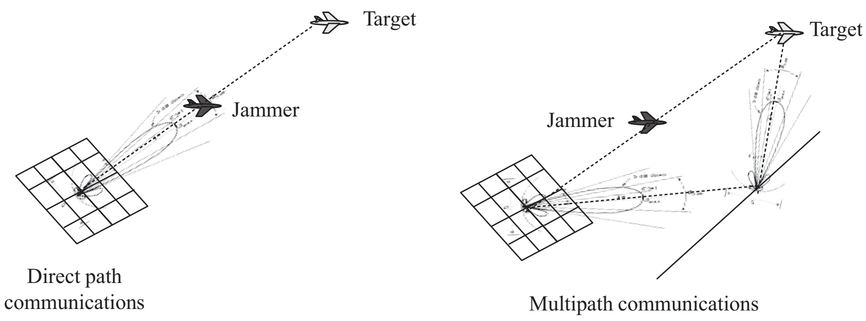


Figure 9.25 Multipath used for communications when direct path is blocked or jammed

or jammers in the main path and automatically steers the AESA to a multipath position to receive the desired signal.

9.2.9 Multiple antennas

Multiple antennas can be used on the base platform, the remote user, or both. The types of multiple antennas are shown in Figure 9.26.

MIMO defines multiple antennas at both the transmitter and the receiver for increased data rates or robustness to the harsh environment including antijam, multipath, and noise performance. Multiple-in, single-out identifies multiple antennas at the transmitter and a single antenna at the receiver for transmitter antenna diversity. The single-in, multiple-out defines a single antenna at the transmitter and multiple antennas at the receiver, which is used for receiver antenna diversity which is primarily used to mitigate multipath.

MIMO is used to increase data rates or improve antijam. The trade-off between data rate and antijam can be adaptive due to the environmental. If the jammer and noise levels are high, MIMO is used for improved antijam performance by sending the same data through multiple antennas using multiple paths. If one of the paths is unreliable due to jamming or multipath, the other paths contain reliable data. If the jammer and noise levels are low, then MIMO is used for improved data rates by sending different data in each of the paths and then combining these data paths in the receiver.

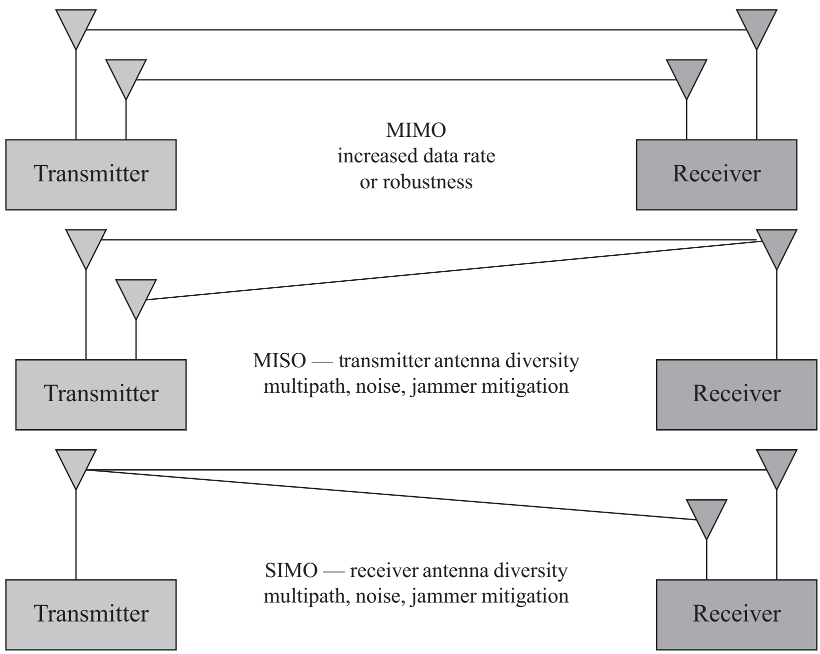


Figure 9.26 Multiple antenna configurations and definitions

In addition, antenna diversity can be used to mitigate multipath (Figure 9.27). This includes using multiple antennas at the receiver end to cancel the unwanted multipath. Since the path lengths are different, if one of the antennas is in a multipath null, then the other one is not in the null since the null changes according to position and distance to the transmitter (Figure 9.27).

9.2.10 Network configurations

Networks can be adaptively configured to provide the optimal solution for a cognitive system. One of the aspects of a network is the ability to do multihop. Multihop networks can communicate to one node of the network through another node, or relay node. For example, if a direct path to a specific remote user is blocked or being jammed, the cognitive network can sense the problem and perform a

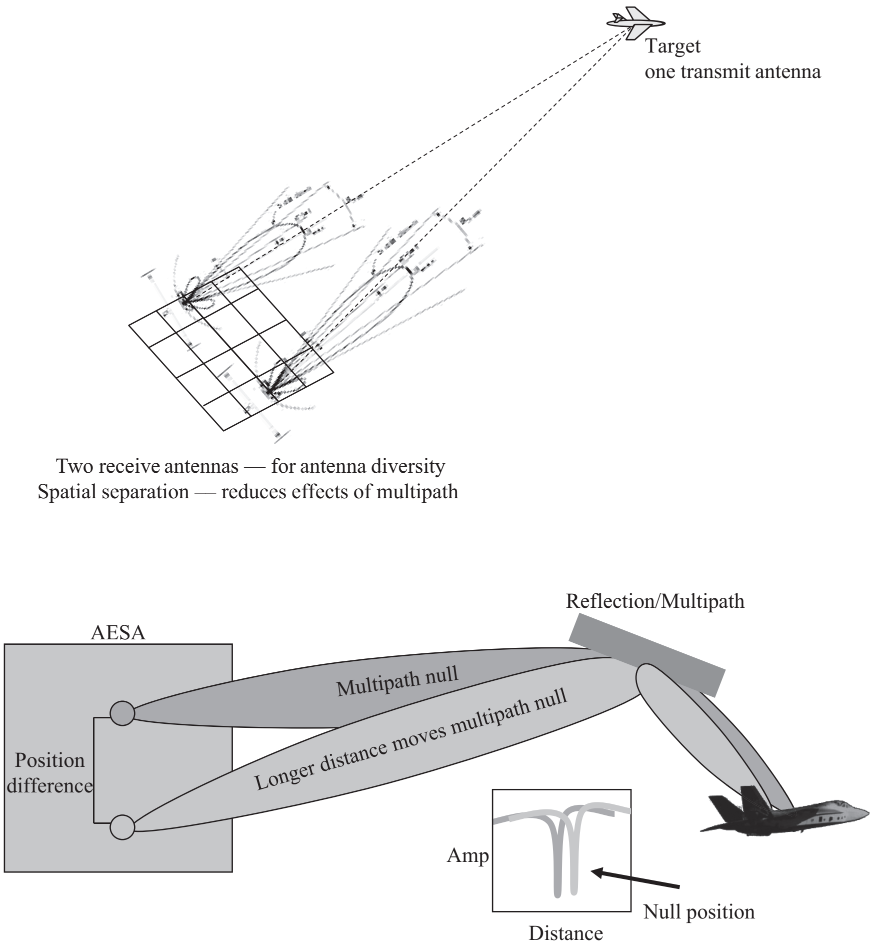


Figure 9.27 Antenna diversity mitigates multipath

multihop to the remote user via another user or node (Figure 9.28). If the system contained power control only for the adaptive process, the power would continue to increase to overcome the obstacle making the users vulnerable for detection and also causing unwanted jamming of other nearby network users or other systems. In the case of blockage, DSA would not be successful since the blockage would attenuate all frequencies. Multipath communications may be an alternative, but it may be difficult to find a clear multipath direction and is actually more complicated than simply using the network configuration to solve the problem. Implementing a cognitive system approach that can evaluate and process the information and select the optimal solution from several options and capabilities would produce the best approach. In this case, the system would evaluate all of the possible solutions and choose the network multihop solution for the optimal performance.

9.2.11 Cognitive MANET networks

Cognitive networks are used to analyze and reconfigure the network according to the changing environment. Many network capabilities can be monitored and made adaptive. Mesh and mobile ad hoc networks (MANETs) have the ability to interconnect multiple nodes either directly or using multihop techniques. The latter uses a node in the network as a relay to send communications to another node, which extends the network's range and also allows the network to adapt to a changing environment which is referred to as range extension.

Since MANETs are constantly changing due to the mobility of the network, they can use adaptive techniques as they self-form and self-heal the ever-changing network and environment. In addition, a cognitive network possesses the ability to learn from the consequences of the cognitive action that was utilized to account for the changing environment. This is important in determining how well the cognitive change affected the overall system or network and using that knowledge to provide a better overall solution. This introduces a smart, learning process that can

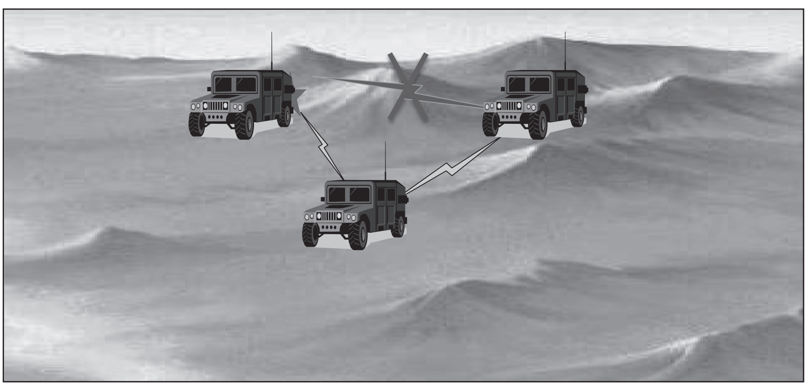


Figure 9.28 Network multihop solution overcomes blockage or jamming of the direct link

continually improve the performance of the network and system. MANETs can monitor and adapt to many changing environmental capabilities, see Table 9.1.

The list in Table 9.1 is extensive and constantly growing with new technology and increased network demands. Many techniques, including gaming theory, are used to determine the best approach for adapting networks for the optimal performance. The network decisions are modeled using all of the capabilities of the individual components to provide a complete cognitive solution. These approaches attempt to provide the optimal solutions given many variables that exist in a MANET. However, the basic concepts are common to all adaptive techniques and cognitive networks.

Another application in developing a cognitive network or system is to prevent interference to other users of the band including primary users. When the industrial, scientific, and medical band became available for digital cellular telephones, strict

Table 9.1 Cognitive possibilities for an adaptive MANET

Capability	Description
Traffic load and flow	Congestion due to multiple users. Schedule operation times, priority users, data rate
Usage with respect to time of day or missions	Schedule users at different times of the day, and according to mission parameters
Nodes coming in and out of the network	Acquisition and awareness of nodes, more or less nodes, more or less bandwidth
MAC layer organization	Adapts the MAC layer for improved-networking performance
Optimize network for minimum hops, best path in the network	Minimize multihops unless needed, provide the shortest best path for communications
Network topology include multihops	Extends the range of the network and prevents blockage & jamming with multiple hops
Network conditions	Monitors network conditions, jammers, noise, blockage and configures the network for optimal performance
Number of hops permitted	Each network adapts to the environment to determine the number of hops permitted, depends on bandwidth and timing
Link quality between nodes	Adapts to link quality by changing a function, frequency, modulation, power, or multihops
Priority messaging and hierarchy	Provides adaptability for priority users to ensure the required bandwidth for them
Connectivity versus coverage	Makes trade-offs between how well the network is connected versus how far the network can extend
TDMA structure	Adapts the timing and TDMA slot assignment for optimal network performance
CSMA and collision avoidance	Senses the carrier to prevent users from jamming each other
Types of communications: broadcast, multicast, unicast	Adapts to the type of communications that is needed in the network given environment and mission

requirements were necessary to avoid interference with its primary users (i.e., operators in those industries). Spread spectrum techniques were used to prevent jamming.

Now, though, cognitive networks and systems can also prevent interference to the primary users and other users by sensing the bands of operation and techniques incorporated by the primary users and rapidly change their cognitive capabilities to prevent interference. Cognitive networks can also learn users' patterns and use this information to further enhance their cognitive capability. For a simple example, if a user is using the band only at a specific time, then the cognitive network can learn that characteristic and not use the band at that time to prevent interference. The cognitive techniques listed in Table 9.1 can be used to prevent interference in accordance to the learned information of the cognitive network.

9.3 Cognitive system solution

Cognitive devices have progressed from traditional relatively static designs to variable capabilities (Figure 9.29). However, for a system to provide a complete, optimal, cognitive solution by evaluating all of the variable devices, the system cognition uses the information of the environment and available devices and makes the optimal cognitive system solution (Figure 9.29). If the individual devices are cognitive without the system solution, the individual controls may not be optimum, that is, the aforementioned multihop network solution. The M&C system is used for all the capabilities in the system and can make the optimal system cognitive choice.

The main cognitive focus has been on DSA, power control, and networking. A complete cognitive system solution would include these and also would monitor all of the environmental parameters and control all of the available hardware, software, and networking capabilities. The individual capabilities of a system are often

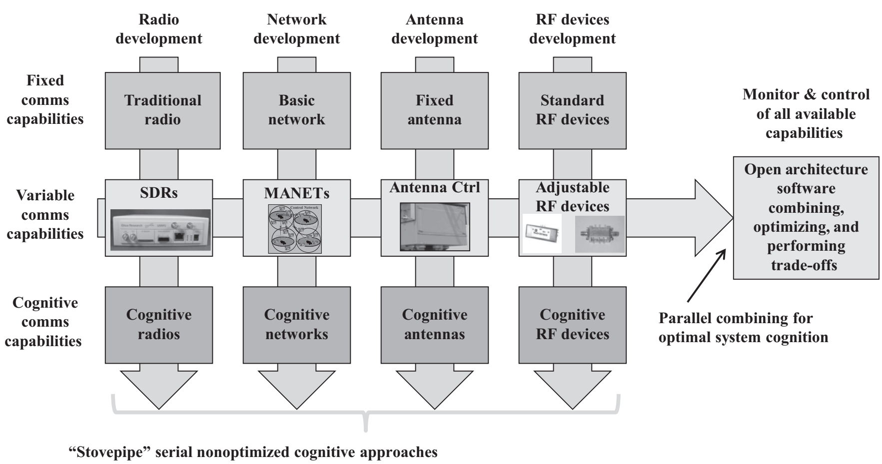


Figure 9.29 Serial versus parallel evolution of cognitive capabilities

times developed in a "stovepipe" approach where each of the capabilities becomes an independent solution to cognition. These separate developments prevent an overall system approach since they are designed to adapt to the environment with their limited capabilities. A much better approach is to utilize all of the variable capabilities of each function with a monitor and control system to make the trade-offs and find the optimal solution which determines which function or functions are controlled and changed according to the environment (Figure 9.29). The approach combines the capabilities of the entire system and makes the optimal decision.

A cognitive system therefore evaluates all of the hardware and software modules that make up the system. The modem contains many aspects of a cognitive system, including DSA, power control, modulation type, error correction, adaptive filters, and spread spectrum. The antenna system also has cognitive capabilities such as beam steering, null steering, MIMO, antenna diversity, beam spoiling, and beam narrowing. AESAs provide the means to accomplish many of these cognitive capabilities. Other hardware such as the RF front ends, transmitter, and receiver can provide Cosite mitigation, multipath mitigation, power control, and DSA. Software and system configurations such as networking topology, trade-offs and optimization, and software monitor and control comprise the rest of the cognition capabilities.

The cognitive system discovers what capabilities are available to the system, evaluates all possible mitigation techniques, determines which solutions are optimal given the known parameters, and implements and controls the hardware and software modules (Figure 9.30). The cognitive system determines the best solution for time, effort, vulnerability, optimization, range, and performance. These decisions are evaluated with respect to a price or cost to the system, which determines the impact to the system's performance before the decision is made. This allows for the optimal solution with the least impact to the system's performance. The system uses a cognitive monitor-and-control (M&C) system that is central to all of the hardware and software modules. This block determines the optimal cognitive solution for the system by monitoring the environment and then controlling the available hardware and software modules.

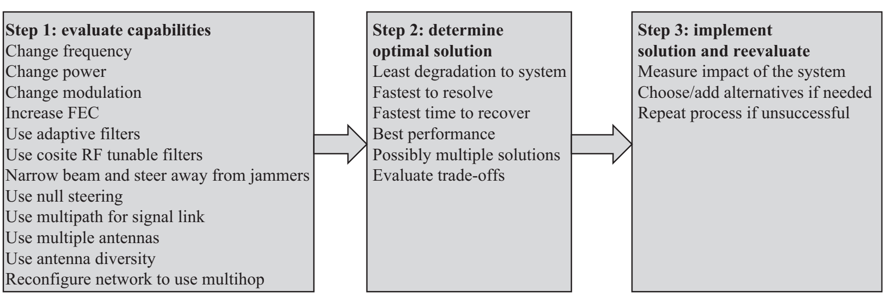


Figure 9.30 Cognitive system processes to provide the optimal solution for the system

A flow diagram shows the process of the cognitive system (Figure 9.31). The incoming energy is monitored and detected and passed on to the next process, which performs the trade-offs. The next process determines the optimal solution and selects the device used (e.g., AESA, power control module), and the final process changes the selected device to mitigate the problem. Once the decision and control has been completed, the results are used for learning and reasoning which is fed back to assist in performing trade-offs and optimization (Figure 9.31).

The performance of any data link system can be reduced by a low-level signal where the SNR is not large enough to prevent a low BER in reference to the noise floor, or it can have a sufficient power level that is being jammed, which produces a low SJR. Each of these scenarios needs to be addressed in a slightly different manner by the cognitive solution. An example is using some of the capabilities of a cognitive system addressing both these needs (Figure 9.32). If the signal is being jammed by a narrowband jammer, an adaptive filter could be enabled to mitigate the jammer. In addition, it could implement further mitigation techniques such as changing the frequency, increasing the power, or changing the modulation, beam steering, or beam nulling. The broadband signal does not utilize the adaptive filter but has other techniques at its disposal. In addition, both narrowband and broadband jammers can be mitigated by reconfiguring the network or using multipath communications. If the system does not have some of these capabilities, they can be skipped, and if they have additional capabilities, they can be employed. When the degradation is due to a low signal level and not jamming, many of the same techniques can be used, but without the adaptive filter or beam steering and nulling. MIMO and antenna diversity would apply, if the signal is in a noisy environment or in a multipath null (Figure 9.32).

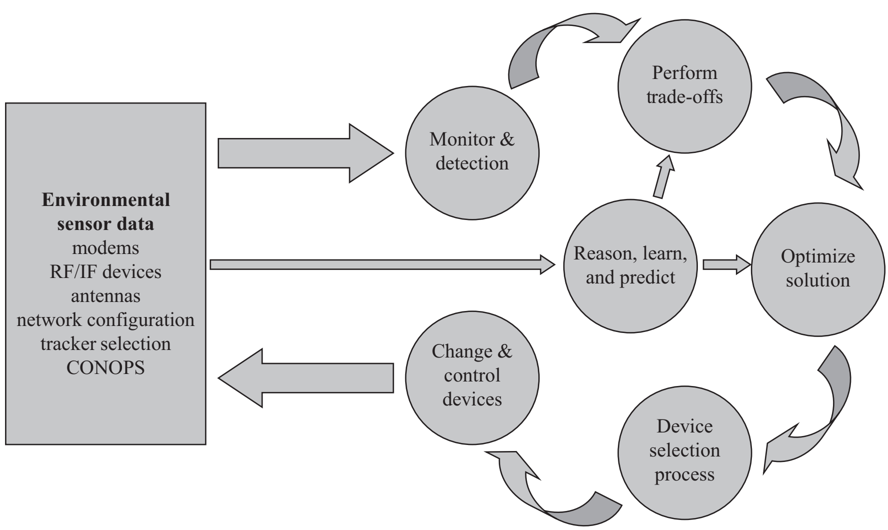


Figure 9.31 Cognitive system flow diagram showing the process for cognitive implementation

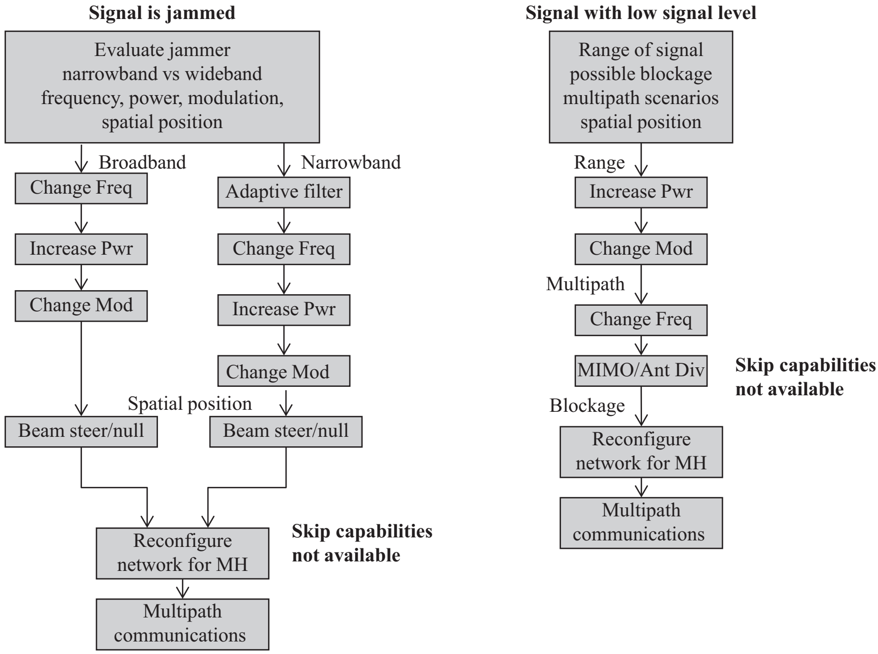


Figure 9.32 Cognitive system addresses both the low SJR and the SNR separately

Since there are many capabilities and system solution options for each of the scenarios, it is not practical to describe them all. The optimal solution will depend on the application. One approach to combining several capabilities is shown in Figure 9.33 for the base platform and in Figure 9.34 for the remote platform. The decision blocks with $C = n$ denote the number of times the change has occurred. For example, if the path shows to change the frequency, if the number of times the frequency has changed in the flow equals to C, then the flow continues to the next change, in this case the change in modulation.

The base platform approach starts with the power control capability by comparing the received RSS to a threshold and raising or lowering the power level, which is verified by the BER or LinkLock. The latter is generally sufficient to determine if it is the correct signal. The numbers in the decision blocks indicate the number of times it changes. It then branches off to either a multihop network or frequency control. If this is not successful, then it changes the modulation type. Beam forming and null steering are next, and if that is unsuccessful, then MIMO is reconfigured and the adaptive filter is enabled (Figure 9.33).

The remote platform approach starts the power control capability by receiving the RSS level and the BER/LinkLock verifier. If the BER/LinkLock is good, then the received RSS level is compared with the previously received RSS level. If it is lower, then the remote decreases its power output, and if it is higher, then the remote increases its power output. If the BER/LinkLock is bad, then it branches off

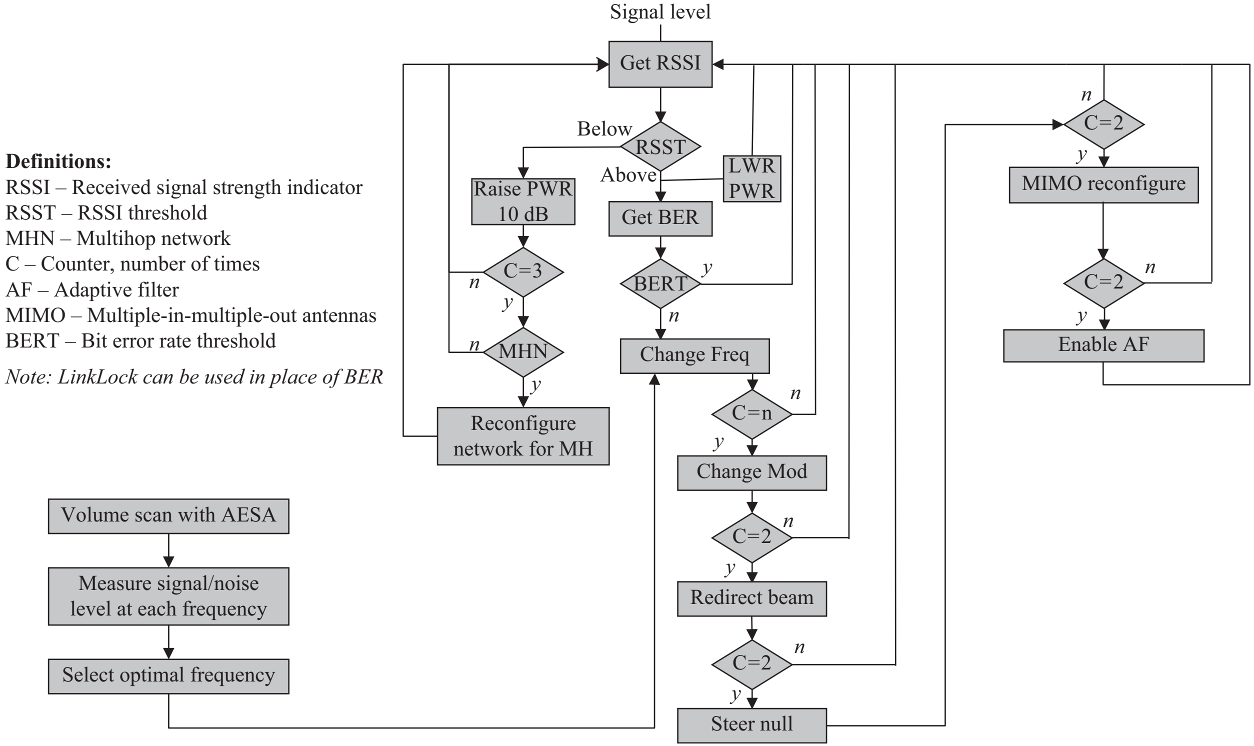
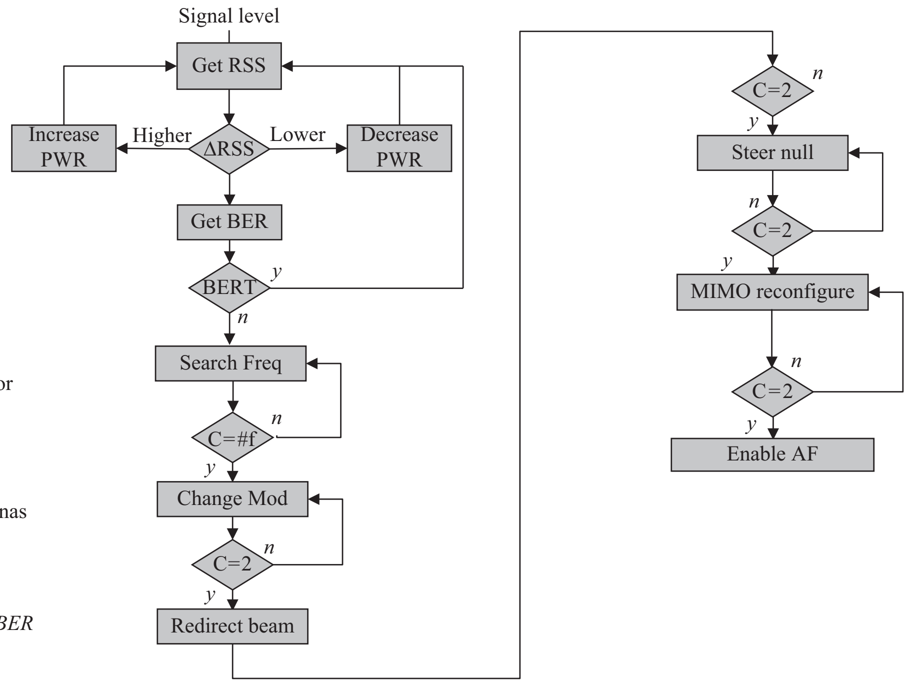


Figure 9.33 Cognitive system of the base platform using multiple cognitive capabilities



Definitions:
RSSI – Received signal strength indicator
RSST – RSSI threshold
MHN – Multihop network
C – Counter, number of times
AF – Adaptive filter
MIMO – Multiple-in-multiple-out antennas
BERT – Bit error rate threshold
ΔRSSI – Delta between measurements

Note: LinkLock can be used in place of BER

Figure 9.34 Cognitive system of the remote platform with multiple cognitive capabilities

to a frequency search to find the correct base frequency. If that is unsuccessful a given number of times, then the modulation is changed, the beam and null steering are enabled, and finally MIMO is used and the adaptive filter is enabled (Figure 9.34).

9.4 Predicted methods—predicting before signal loss

Prediction is another key element and advantage in developing a cognitive solution. Predicting signal losses before they occur allows the system to reconfigure using cognitive processes before the signal is lost (Figure 9.35). Once the signal is lost, then the system needs to acquire the signal again which takes some time and processing. The cognitive system monitors the dynamics of the system and determines the out-of-range time when the prediction is evaluated. The step solution is shown below:

1. Establish current position (range)
2. Perform a link budget analysis to determine maximum range with margin
3. Calculate relative velocity
4. Calculate the predicted time when the signal is at maximum range

The system then makes the necessary adjustments before the signal loss using the following options:

1. Increase power
2. Lower data rate
3. Decrease overhead functions
4. Narrow the beam of the antenna
5. Reconfigure the network for multihop

Any one of these solutions or combination of these solutions can be utilized. The system controller determines the optimal solution with minimum cost to the system.

The range predicted outages for a single user predicts the position using the following:

1. Use closed-loop α/β tracker position prediction
2. Take the difference between two positions
3. Measure the time between the two positions
4. Calculate the velocity
5. Multiply velocity by the time for the next position
6. Predict the next position
7. Calculate next position's range
8. Compare range with link margin from the link budget

The results are shown in Figure 9.35.

The range predicted outages for two users using the following:

1. Calculate range taking the difference of their positions using received Nav/GPS data

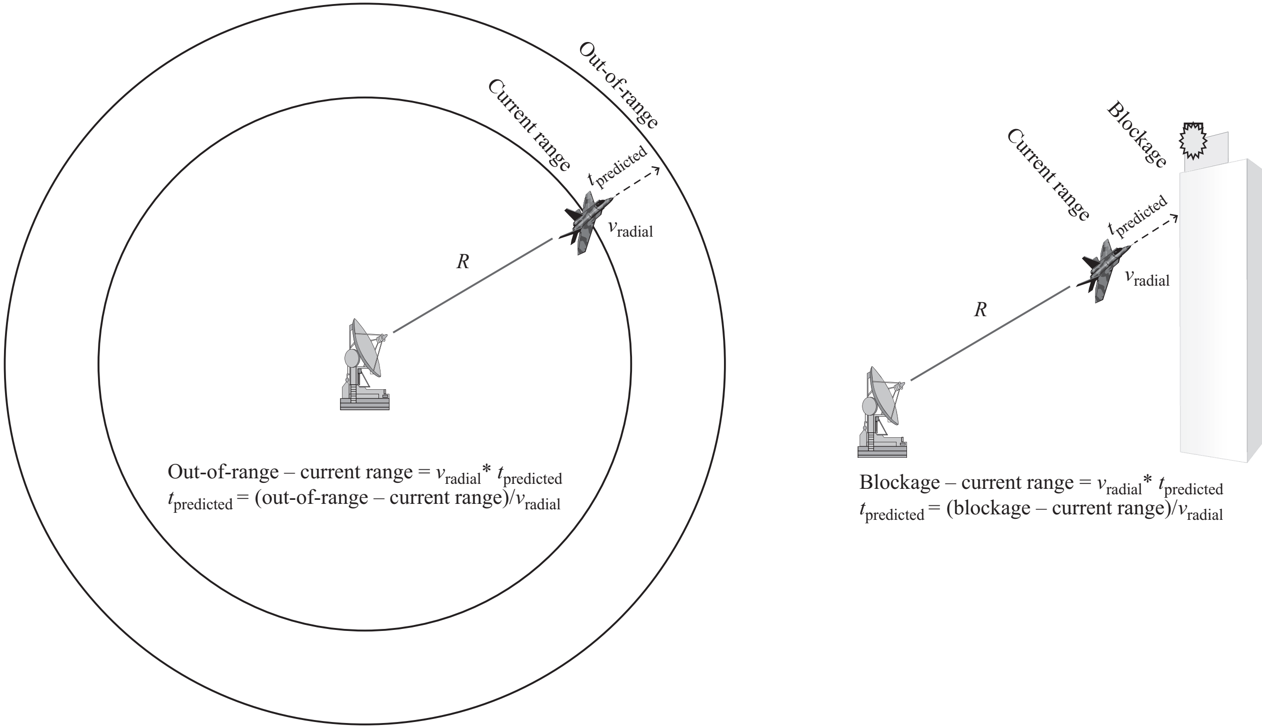


Figure 9.35   Cognitive capability predicts time to signal loss

2. Link budget—set threshold for switching at a given range, use <10 dB link margin (adjustable)
3. Predicted position and range using relative velocity between users

Predicting outages due to blockage and/or jammers uses the following:

1. Monitor and determine the time for expected blockage or jammers
2. Establish current position (range)
3. Monitor and determine position of blockage or jammers
4. Calculate relative velocity
5. Calculate the predicted time when the signal is blocked or jammed
6. Determine relative positions of both antennas
7. Measure roll, pitch, and yaw of both aircraft
8. Make adjustments before signal loss

The results are shown in Figure 9.35.

Other factors used in cognitive system for predicting signal degradation as follows:

1. The time of day
   - (i) Band usage—Certain times during the day there are more users predicted to use the bandwidth which causes more interference in the desired cognitive system (Figure 9.36). Predict the band usage, provide more antijam/spread spectrum, lower the data rates, and increase power.
   - (ii) Ionosphere effects—Change during the time of day which includes losses and delays. Cognitive systems adapt the power needed and compensate for the additional time delay.
2. Atmosphere losses such as rain, snow, moisture
   - (i) Monitoring the time of day, looking at weather reports and other monitoring functions for atmospheric losses.

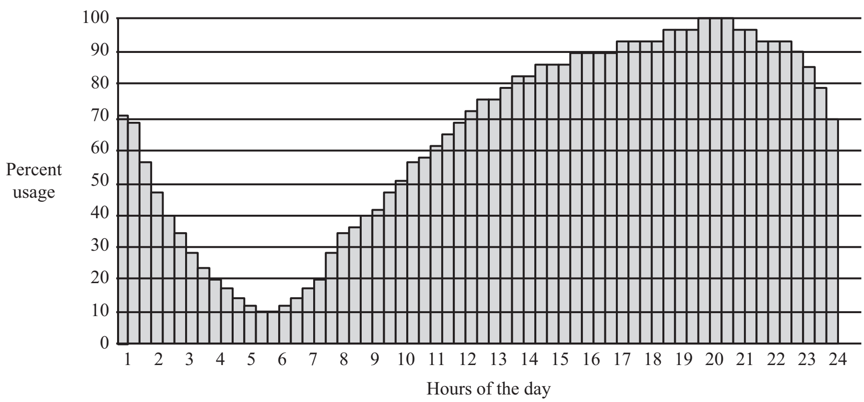


Figure 9.36 Smart phone usage during hours of the day

(ii) The cognitive system can change the power, data rate, and other factors to predict when changes need to be made.

9.5 Learning and reasoning capabilities of a cognitive system

A cognitive system can learn from past decisions to influence future behavior. This system improves through experience gained over a period of time, and the results are retained in memory. This process is used for similar circumstances in the future and also provides a starting point or what to avoid.

The learning process is used as an input for improved reasoning performance, it provides higher accuracy of the model which improves as a result of learning, and behavioral patterns are established by using observations to update a deterministic or stochastic model of the environment.

Reasoning is an immediate decision process that must be made using available historical knowledge. This knowledge is about the current state of the system. The system gathers relevant information from the knowledge base and sensory inputs (observations) and decides on a set of actions. Learning evaluates relationships between past actions and current observations and also between different concurrent observations. The system then converts this to knowledge to be stored in the knowledge base. This is usually a long-term process (could also be short-term for immediate decision-making) and is an accumulation of knowledge based on the perceived results of past actions. It provides emphasis on enabling reconfiguration at all layers of the protocol stack. Reconfiguration can vary from changing parameters within a layer or the replacement of the entire layer with a different protocol.

A standard agent model performs observations, actions using an inference engine and a knowledge base. The knowledge base stores the information, and the inference engine applies logical rules to the knowledge base and deduces new knowledge. Therefore, reasoning and learning are combined with the operation of the inference engine and knowledge base.

This model works with the cognitive architectures previously mentioned.

9.5.1 Methods of reasoning

There are different algorithms and techniques that are used in learning and reasoning. One of the methods is known as a heuristic reasoning method. A heuristic method uses experience-based techniques for problem solving, learning, and discovery which gives a solution which is not guaranteed to be optimal. It is also known as a rule-of-thumb, educated guess, an intuitive judgment, stereotyping, or just common sense. It uses approximation algorithms which speed up the process of finding a satisfactory solution via mental shortcuts to ease the cognitive load of making a decision.

General reasoning method provides a detailed algorithm and takes the time to optimize the best solution for the problems and is generally better since heuristic methods are limited in the range of problems to address. The trade-off is the time it

takes to arrive to the optimal solution. Often times, the heuristic reasoning method is used when it needs a short time for the solution and can operate in a less than ideal solution.

9.5.2 Elements of reasoning

The purpose of reasoning is to select actions in response to perceived network conditions including historical knowledge and current observations of the network's state. There are two types of reasoning, which are as follows:

1. Deductive—forgoes hypotheses, conclusions based on logical connections.
2. Inductive—forms hypotheses that seem likely based on detected patterns. This is generally the preferred for communication systems.

A single decision is an action based on immediately available information. This is good when immediate action is required, for example, a quick response to congestion in a network.

A sequential decision is based on multiple intermediate actions. The process selects the intermediate actions and observes the response of the system following each action. Each intermediate action narrows the solution space until the final action is selected. This is useful for proactive reasoning where time constraints are more relaxed, and there are only indications of an impending problem.

9.6 Multiagent system MAS

There are three concepts when discussing a multiagent system. These are situated, autonomous, and flexible.

The situated concept uses agents that are capable of sensing and acting upon their environment. The agents have incomplete knowledge, partial control of the environment, or both.

The autonomous concept uses agents that have the freedom to act independently of humans or other agents. They may have possible constraints on the degree of autonomy each agent has.

The flexible concept uses agents where their responses to environmental changes are timely and proactive. These agents interact with each other and possibly humans to solve problems and assist other agents providing the responses.

Cooperative distributed problem solving is an inherent group coherence which provides motivation for the agents to work together and is inherent in the system design.

9.7 Noncooperative game theory

Each capability maximizes their own solutions and do not care about other nodes in the network are considered to be selfish nodes. For example, power control for a selfish node increases power for poor signal without considering that increased

power may jam other users. A selfish node does not care about others in the network. A more ideal system solution may be to lower power and configure the network for a multihop. With the noncooperative game theory, each capability is focused on maximizing their own functions, for example, maximizing their data rate with their own capabilities and not concern with the network and system capabilities.

9.8 Cooperative game theory

Each of the nodes operates in a cooperative behavior using a central processor. The central processor has the ability to integrate all the capabilities available and produce an optimal network or system solution. The cooperative game theory focuses on the total network or system needs and provides the optimal solution using each of the capabilities available. This process makes trade-offs between capabilities for optimal network solution using interactive strategic decision-making for a cooperative solution. Software models are generated to determine the stable operation points for networks. The cooperative network requires self-organizing, decentralized, autonomous flexible mobile networks. They provide multiple access networks for multiple users where each user tries to enter the network by transmitting at the same time. This can cause collisions which is resolved by collision avoidance where each of the nodes that are trying to transmit at the same time are assigned to a different delay depending on a pseudorandom sequence so that the probability that they access the network at the same time is unlikely.

9.9 Coalitional game theory

Coalitional game theory is used for modeling cooperative behavior and is used in forming cooperative groups or networks. The concept of this theory is based on the principle that cooperation strengthens the network or system's performance. It is a powerful tool for modeling cooperative behavior for cognitive networks and Mobile Ad hoc Networks MANETs. There are several applications that utilize this theory including SATCOM On-The-Move SOTM systems, MIMO systems, multimedia applications, machine-learning applications, and resource management and security. This is a powerful tool to study these network scenarios.

9.10 Nash equilibrium

All cooperative types of networks strive to establish a "NASH equilibrium" for optimal system results. It begins with the principle that each of the users knows the equilibrium strategies of other users. Each of the users makes changes taking into account other user's strategies. Since it is cooperative, a single user does not gain doing just their own strategy. All users make the best decisions based on the other users in the network. It makes the best response of each user strategy to all other strategies in equilibrium.

9.10.1 Nash equilibrium definitions

Here is a list of definitions for a system to have a Nash equilibrium:

1. If all strategies are known, can a user improve or benefit by changing strategy?
   (i) If yes, then it is not a Nash equilibrium.
   (ii) If no or indifferent, then the set of strategies is a Nash equilibrium.
2. Pareto optimal—it is impossible to make one user better off without making at least one individual worse off.
3. Pareto improvement—makes one user better off without making any other user worse off.
4. Pareto optimal front—One goal cannot be optimized without affecting another.
5. Cognitive-specific Language CSL—translates end-to-end goals to local element goals.

9.10.2 Nash equilibrium examples

Example 1: A user wants to change frequency to improve its strategies.
How does this affect the other user's strategies in the network if it changes frequency?
How much time does it take to reconnect? Does it affect other user's strategies?
Are there users that have a Cosite jammer with the same frequency?
Example 2: A user wants to reconfigure the network to improve its strategies.
How does that affect current network protocols of other users?
How much time to reconfigure the network?
What multihops are already in place that might be affected?
Example 3: A user wants to increase the power for better SNR.
Does the higher power jam adjacent users?
Does it affect the near/far problem for other users?

9.11 Challenges to cognition

One of the biggest challenges in implementing cognitive systems due to the fact that they are always adapting to different parameters are the rules and regulations that govern the use of the space. For example, how does the FCC specify these parameters when they are constantly changing and how do they enforce them?

One of the approaches is to put limits on the amount of cognition and changes that a system can incorporate. Putting limits on prevents the cognitive systems from getting out of control, or malfunctions and moves to an inappropriate state. However, this can limit the reason to use cognitive systems in the first place.

There needs to be a focus on fail safe implementations which could possibly cause an unsafe environment.

In addition, the interaction between multiple cognitive systems could also be a challenge, for example, ways to prevent collision avoidance or independent users coming to the same conclusions.

9.12 Summary

Cognitive systems are used to adapt the operating system to the changing environment. This can be accomplished by many methods suggested in this chapter, and many additional methods will be available in the future. The optimal cognitive system solution evaluates a system's available capabilities as well as all of the available knowledge about the changing environment, and then it calculates, makes trade-offs, and determines the best solution for the system to adapt to these environmental changes with minimal impact to performance. In addition, the cognitive system contains a learning capability that uses past experiences and impact/results of changes and uses this information to make smart decisions in the future. Game theory has been used to establish cognitive networks, and selfish nodes are eliminated by cooperation and establishing a Nash equilibrium. Although there are challenges with cognitive systems, they are beginning to infiltrate our current technologies, and they produce optimal communications and network systems.

9.13 Problems

1. What does cognitive mean?
2. What are the three main cognitive capabilities of a cognitive radio?
3. What radio technology has been instrumental in implementing cognitive radios?
4. What does DSA stand for and how is it used to mitigate problems?
5. Describe an alternate DSA process using a known random sequence.
6. What is adaptive power control? How does it work between a base and remote?
7. What is the difference between a closed-loop and open-loop design?
8. Who sets the threshold for power control, the base, the remote, or both?
9. What does the remote depend on to adjust its power?
10. How does the addition of the remote device change the control loop?
11. What is another method of power control that is used in cell phones?
12. What are the two basic trade-offs when selecting the type of modulation?
13. Describe one technique on how an AESA can be used to mitigate interference.
14. What two advantages are provided by using MIMO antennas?
15. What does a network use to extend range and eliminate direct path jamming?
16. What multipath advantage method can be used when direct path is unavailable?
17. What does MANET stand for?
18. What is the first layer in the seven-layer comms model and what does it cover?
19. Describe a scenario where a signal is blocked and a cognitive network solution would be better than a cognitive radio solution.
20. What two functions can a cognitive system utilize for future occurrences?

21. What theory is used to work out scenarios in networks and evaluate their performance?
22. Describe what Nash equilibrium means in a game solution or network.
23. What does M&C stand for and why are they important to a cognitive system?
24. Describe why a cognitive system provides a better cognitive solution over individual cognitive solutions?
25. What is one of the biggest challenges for a cognitive system?

Further reading

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Chapter 10

Volume search, acquisition, and track

Wireless communication systems that contain users that are moving in position and incorporate directional antennas require volume search, acquisition, and tracking methods to discover and maintain the communication link. Volume search is used to discover the user, acquisition is used to acquire the user by narrowing down the position of the user in order to point the antenna in the direction of the user, and tracking is used to lock on to the user's position and maintains the position as the user is moving in order to point the antenna in the direction of the user.

There are several methods used to implement volume search, acquisition, and track depending on the requirements and capabilities of each system.

10.1 Volume search

The first step in minimizing the time to search a volume of space is to reduce the search window, or the area that needs to be searched. For example, if both the users are on the ground, then the search window can be reduced by eliminating the elevation search and only look for users in the horizontal search. Also, if a shipboard user is looking for an aircraft user, then the elevation search window would be reduced to only look for users from the 0° to 90°. Windowing incorporates the practical limitations of the system to reduce volume search.

Other ways to reduce the search window is using previous system information, such as the previous tracking information or last known position of the user with time. The window is reduced by predicting the position of the user by using the velocity over the time elapse to estimate the maximum distance the user could have traveled to reduce the search window for that user.

Another way to reduce the search window is increasing the area that the directional antenna can cover. For example, in the case of Active Electronic Scanned/Steered Array (AESA) antennas, the beam can be spoiled or widened to cover more volume. This generates a trade-off between greater coverage and reducing the gain of the antenna or the range of the data link. In addition, the sidelobes of the antenna, if they are not too far down in gain from the mainlobe, can be used to reduce the search window for close-in users.

If other equipment is available such as GPS information that is being received using an omnidirectional antenna, the GPS position can be used as a starting point to reduce the search time.

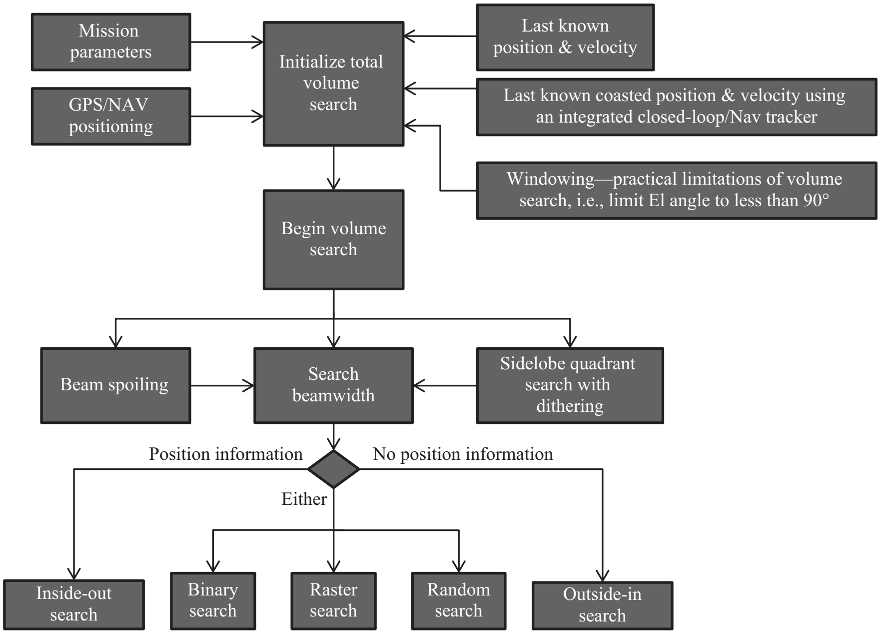


Figure 10.1 Directional volume search process

In addition, any knowledge of the expected user's position will help to reduce the search window, such as Mission parameters specifying the area the user is expected to be, basic direction that the user will be such as from the North, or in front of the antenna and not behind, and many other examples.

There are several techniques to produce volume search for finding an unknown user. The type of search method depends on the time required to find the user, previous knowledge of the user, and the guarantee that the user will not move out of the search window before it is found. A flow diagram shows the various approaches to narrow the search and to determine the optimal search engine (Figure 10.1). These methods of search will be discussed in detail.

10.1.1 Raster scan

Raster scan is one of the simplest ways to scan the search window. It starts at the top left corner and scans horizontally across the search window until it reaches the end, moves down in elevation or row, and scans horizontally until it reaches the end. This process is continued until all of the possible elevation points are completed for the given search window, see Figure 10.2. This is similar to a standard television set.

The scan stops when it receives a valid signal level from a user, which if it is the desired user, there is generally a lock detect indication that it is the correct user. Once the user is found, the system switches to acquisition mode where it does a fine search or acquisition around a small area to verify that it was a valid user. If it turns

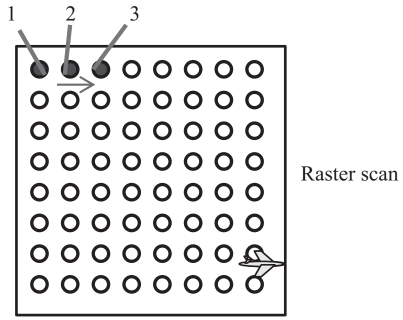


Figure 10.2 Simple raster scan volume search

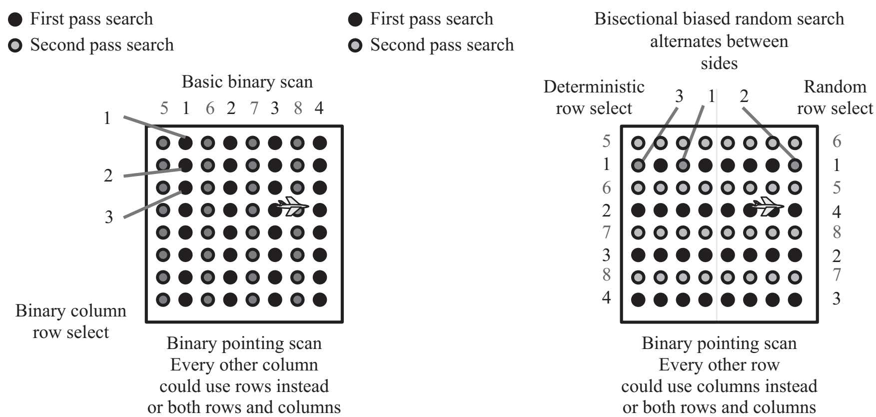


Figure 10.3 Binary scan volume search with deterministic or random bias

out to be a false alarm, or the lock detect did not indicate the correct user, then the raster scan continues from the point of the false alarm.

This is the least desired scan to find a user in a search window; however, it is used as a comparison for other types of improved scanning techniques. The reason that this is the lowest performer is the fact that the user could move out of the area before it finds the user. It is a simple way to cover the search window and can be used for slow-moving users.

10.1.2 Binary scan

Binary scan is formed by scanning every other row, column, or both row and column of the volume, see Figure 10.3. Each column in the basic binary scan example starts with column 1 and scans all of the points in column one, then starts the scan on column 2. The binary scan skips every other column, see Figure 10.3.

In order to improve the basic binary scan, a bisectional bias can be incorporated so that each consecutive scan point is forced to be in the other half of the search window. In addition, the selection can be either deterministic where each of the points in a row are sequential, or it can be a random selection of the points in the row with the bisectional bias as shown in Figure 10.3. For the random bisectional biased operation, the points are still forced to alternate between the two halves of the search window, but the points in each of the halves can be randomly selected. This improves the performance of finding the user in less time.

10.1.3 Random scan

Random scan search uses a pseudorandom pattern for selecting the search points, see Figure 10.4. All random searches are set to not repeat any search point until all of the points have been searched, at which time the random search starts over again.

In order to improve the random search pattern and prevent the system from searching a small area according to the pseudorandom code, a bisectional bias is implemented to force every other search point to be in a different section of the search window, see Figure 10.4. To further enhance the performance of the random search pattern, the search window is further broken down into four quadrants, where each random selected point is forced to be in a different quadrant. Therefore, each of the four consecutive points of the random search will be in a different quadrant, see Figure 10.4. This prevents the random search pattern from searching too long in one bisectional or quadrant, which improves the discovery time for finding a user in the search window.

10.1.4 Bias expansion scan

The bias expansion scan increases the divisions from quadrant to multiple cells. It divides up the search window into smaller search cells, see Figure 10.5. The cell size is variable with the smaller cell size approaching the random search with no bias, and the larger cell approaches the quadrant bias. The optimal cell size depends on several factors including search time, resolution, search window size, and overall system requirements.

The example shows a selection of 16 cells with five search positions in each cell, see Figure 10.5. A cell is randomly selected and a search position is randomly

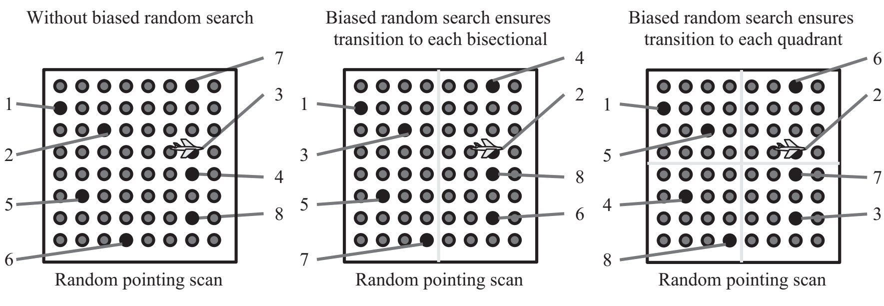


Figure 10.4 Random volume search with sector bias

searched. Every cell is selected once through the random selection process to prevent repeating a cell. Each position is searched once through the random search process to prevent repeating a search position. Cells are searched again with random offset if the user is not found. This provides the optimal volume search with respect to finding a user in a search window compared to the other types of volume-search techniques.

10.1.5 Outside-in volume search

In order to optimize a volume-search technique that is designed to prevent losing a user during the volume search, an outside-in biased random search is used if the location of the user is unknown with no past position information available.

This technique uses a biased random search on the outside perimeter and works its way into the center of the search window, one row at a time, see Figure 10.6. Multiple rows can be combined to increase the search time with a higher probability of losing the user. Each perimeter search points are randomly selected until all of the points have been searched for that perimeter, then the search moves to the next inside perimeter and repeats the process until it has searched the total volume defined. Bisectional and quadrant biases can be applied as before to optimize the search, see Figure 10.6. In addition, the search window can be expanded to meet the total volume search if needed.

10.1.6 Inside-out volume search

This search technique is used when the previous positioning information of the user is available. It searches the last known or predicted location and moves outwards. It can also use external sources such as GPS or other navigation position information that is available including the last known or predicted closed-loop track information or last known or predicted internal navigation position data that was sent through the data link.

This search method uses the same techniques as the other search solutions such as biased random search on the perimeter but starting at the inside perimeter at the predicted location of the user and working its way outwards, see Figure 10.7.

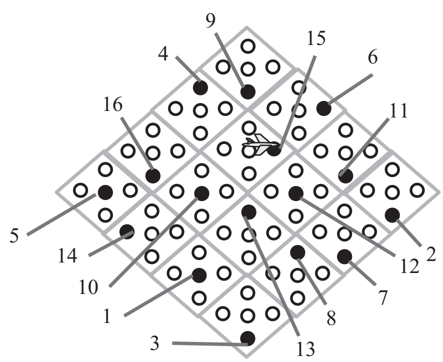


Figure 10.5 Sector bias expansion for optimal volume search

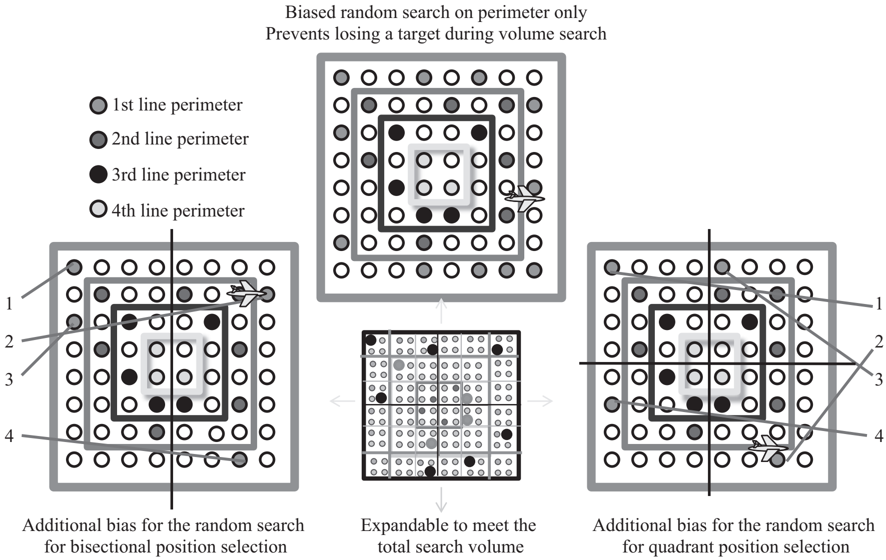


Figure 10.6 Outside-in volume search with no positioning information

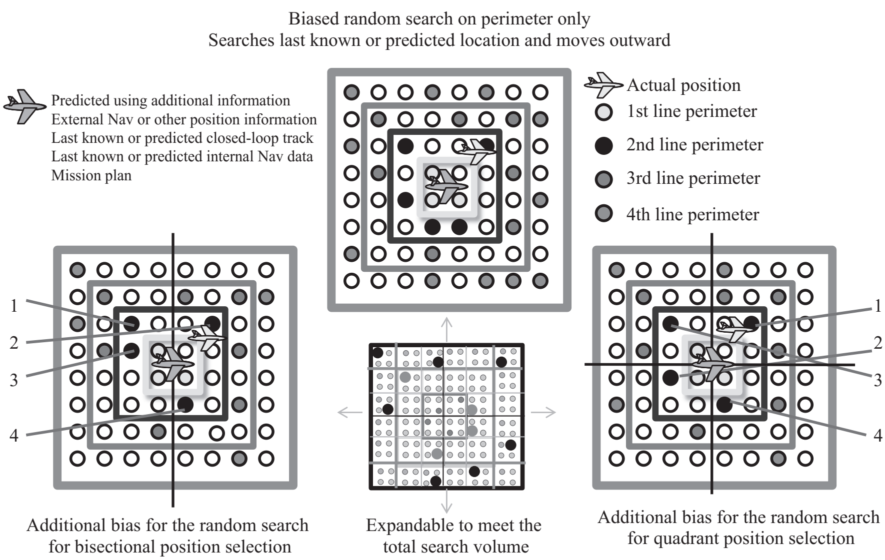


Figure 10.7 Inside-out volume search for previous positioning information

Table 10.1 Time and missed detection performance search results

Time performance results	Missed detection performance
1. Random with bias expansion	1. Random single perimeter outside-in with quadrant bias
2. Random quadrant bias	2. Random double perimeter outside-in with quadrant bias
3. Random bisectional bias	3. Random single perimeter outside-in with bisectional bias
4. Random	4. Random double perimeter outside-in with bisectional bias
5. Random double perimeter outside-in	5. Random single perimeter outside-in
6. Binary random row selection with bisectional bias	6. Random double perimeter outside-in
7. Random single perimeter outside-in	7. Random quadrant bias
	8. Random bisectional bias
	9. Random with bin multiplier
	10. Random
	11. Binary random row selection with bisectional bias
	12. Binary with bisectional bias
	13. Binary

In addition, both the bisectional and quadrant biases can be used, and it can be expanded to meet the total volume search as needed.

10.1.7 Directional volume search results

The time and missed detection results are shown in Table 10.1. For the best time performance, the random search with bias is the best approach, with the random with bias extension being the optimum.

The best missed detection performance is the random search using single perimeter outside-in with quadrant bias, see Table 10.1.

The inside-out search is the best performer to reacquire a target with position information known or predicted.

10.1.8 Beam spoiling

Spoiling the antenna beam pattern widens the beamwidth to reduce search volume and search time, see Figure 10.8. The wider beamwidth reduces the time to search; however, the antenna gain is decreased which reduces the range of detection. When searching for a user, the errors can be high as long as there is enough information received to determine if it is a valid user. Most data links require a very low probability of error to receive data such as 10^{-6}; however, for detecting a signal, the probability of error can be much higher, for example 10^{-2}. Once the user is discovered, the beam spoiling can be disabled and the data probability of error is

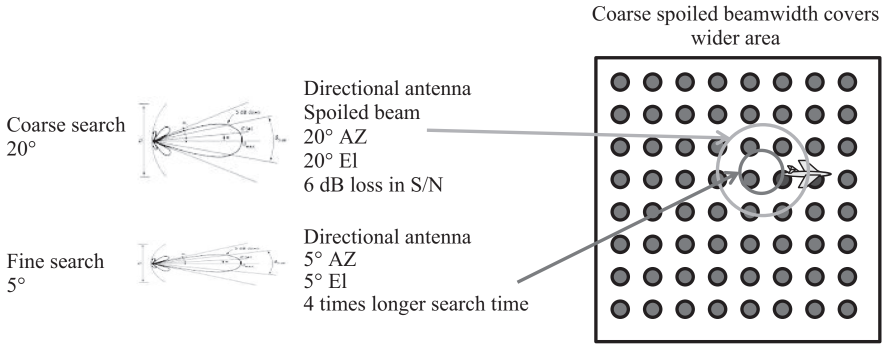


Figure 10.8 Beam spoiling to reduce search volume

back to the lower value for good communications. The following example is shown:

> Example: Spoil the beam by 4 times which reduces the search time by 1/4.
> E_b/N_o is reduced by 6 dB.
> BPSK Data requires 10.5 dB E_b/N_o for Pe $= 10^{-6}$.
> Detection signal is therefore 4.5 dB E_b/N_o which is less than 10^{-2}.

Therefore, spoil the beam to decrease search volume for detection, unspoil the beam once the user is located to provide the required signal level and range. In addition, a spoiled beam could be used as a coarse search, and the unspoiled beam could be used for the fine search for multiple applications. AESA technology can easily and very quickly spoil and unspoil the antenna beam for these types of applications.

The range to the user can be used to determine the amount of beam spoiling required. This can be based on internal or external navigational data using position information or signal strength. If the range is short, then more beam spoiling can be used. If the range is long, then less beam spoiling is needed. Therefore, close-in users can use a much wider beam for much faster time for discovery.

10.1.9 Sidelobe detection

Sidelobes of the antenna can be used to search more area which reduces the search time of the system. Sidelobes are generally down in amplitude; however, similar to beam spoiling, they can still detect signal levels with a reduced gain. In order to use the sidelobes for detection, the magnitude of the sidelobes needs to be determined. Sidelobes can be used for detection, and once the user is found, the mainlobe is pointed toward the user. Quadrant search is used for the sidelobe probability of detect. Each quadrant is broken down to finer levels across the entire antenna face. The antenna face is divided into quadrants, which can then be subdivided again, see Figure 10.9. The search pattern can then be set to account for the sidelobe detection.

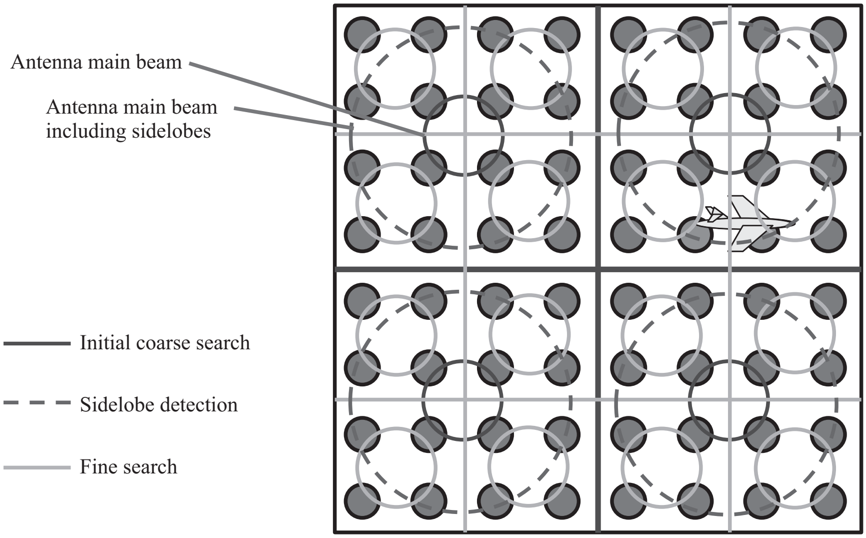


Figure 10.9 Sidelobe detection and dithering to reduce search volume

Another aspect of using sidelobes for detection is the nulls that are present between the sidelobes and the mainlobe. In order to fully utilize the sidelobes, the antenna pattern uses dithering of the beam during search to prevent signal loss between the mainlobe and sidelobes. AESAs have the ability to provide this technique.

An example of using the sidelobes is shown below:

BPSK requires $E_b/N_o = 10.5$ for $\mathrm{Pe} = 10^{-6}$
1st sidelobe levels = 13 dB down (17 dB for circular)
Sidelobe $E_b/N_o = -2.5$ dB
If $E_b/N_o = 3.5$ dB for detection, then E_b/N_o needs to be increased by 6 dB
Sidelobe detection would work at 1/2 range of the actual user (only for short-range users)

10.1.10 Coordinate conversions for volume-search process

There are two possibilities to provide the pointing directions (Az, El) to the antenna for the search processes. One is to use coordinate conversion for a moving platform that is not stabilized. The pointing directions (Az, El) are referenced to the horizontal plane regardless of the orientation of the moving platforms that can roll, pitch, and yaw referenced to the horizontal plane, for example, a ship on the ocean, see Figure 10.10. The pointing directions to the antenna are reference to the rotated position. Another possibility uses a stabilized fixed platform (no roll, pitch, and yaw) where coordinate conversions are not required.

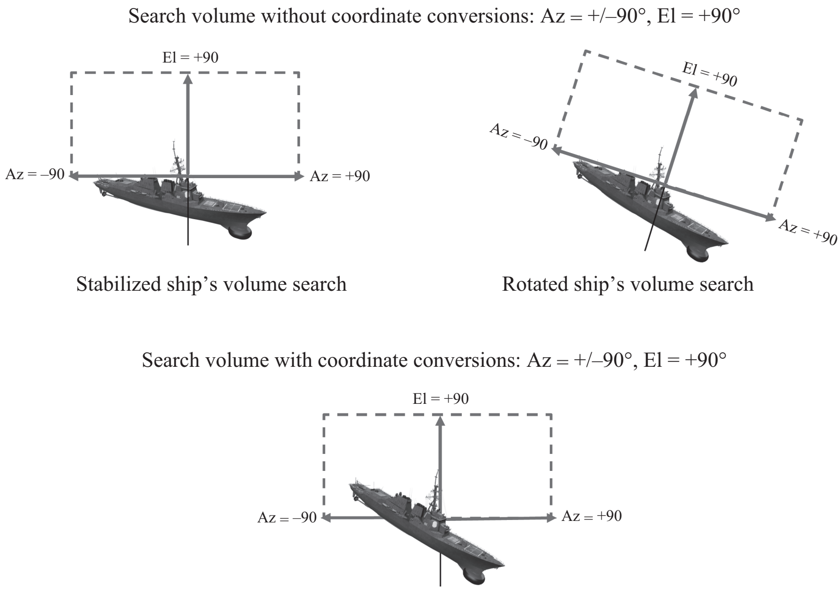


Figure 10.10 Coordinate conversions for nonstabilized platforms

10.1.11 Directional antenna beam

Directional antennas generate a directional beam that focuses the energy in a specified direction. Parabolic antennas are directional beams using a parabola-shaped beam. The diameter of the parabolic antenna determines the beamwidth: the larger the diameter, the narrower the beamwidth. The approximate beamwidth for a parabolic antenna is the following:

$$\theta = \sin^{-1}(\lambda/D)$$

where θ is the 3 dB beamwidth, λ is the wavelength, and D is the diameter of the parabolic antenna.

Other factors that reduce the gain of the antenna are the shape of the antenna, anomalies on the surface, and radomes if used that cover the parabolic antenna.

The following example shows how to calculate the beamwidth:

Given: $f = 2.4$ GHz, $D = 3$ m
$\lambda = C/f = (0.3 \times 10^9\ \text{m/s})/(2.4\ \text{GHz}) = 0.125$ m
$\theta = \sin^{-1}(\lambda/D) = \sin^{-1}((0.125\ \text{m})/(3\ \text{m})) = 2.3°$

10.2 Acquisition: two-dimensional sequential scanning

Once the user has been found during volume search, the user then needs to be acquired, which is a fine search process to establish a link between the users and to narrow the search window in order to lock onto the user and track the user.

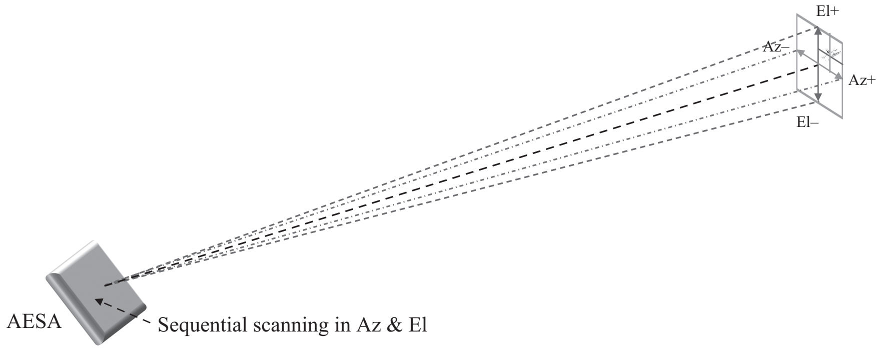


Figure 10.11 Two-dimensional sequential scanning

Sequential scanning of both the azimuth (Az) angle and the elevation (El) angle is used to refine the user's position in order to begin tracking. The sequential scanner scans in +/−El and then +/−Az and measures the received signal strength (RSS). The measured RSS moves the antenna in the direction of the largest received signal in both El and Az, see Figure 10.11. Once the antenna is pointed toward the user with the maximum RSS in both Az and El, the combined angles determine platform position for the system to begin tracking the user.

The Az and El direction information updates the AESA-pointing vector to point toward the user. In addition, the system calculates and estimates the processing delays to reduce lag between antenna scan command and measured RSS. Time stamping all of the data ensures that the RSS matches the antenna position to the user.

A Link Lock indicator is used to ensure that the RSS is coming from a valid user and not from a jammer or unknown source. This Link Lock indicator processes the user's code or PN-sequence to determine the user identification. The user is now acquired and transfers to the track state.

10.3 Track

Track is used to lock onto a user in motion and maintains continuous communications using directional antennas. As the user moves, the antenna receives error signals which are used as feedback signals for pointing the antenna toward the user platform. Feedback keeps the antenna beam pointing at the user. If the signal is lost, the track system predicts the position of the moving user platform and will coast between the received signals. These horizontal and vertical angle errors are used to point the antenna in the correct position in addition to the coasted predicted positions of the moving user platform.

Mechanically steered antennas use gimbals to adjust or rotate the antennas in both Az and El to provide the position. AESAs use electronic steering which is very fast and is able to track much higher dynamic user platforms.

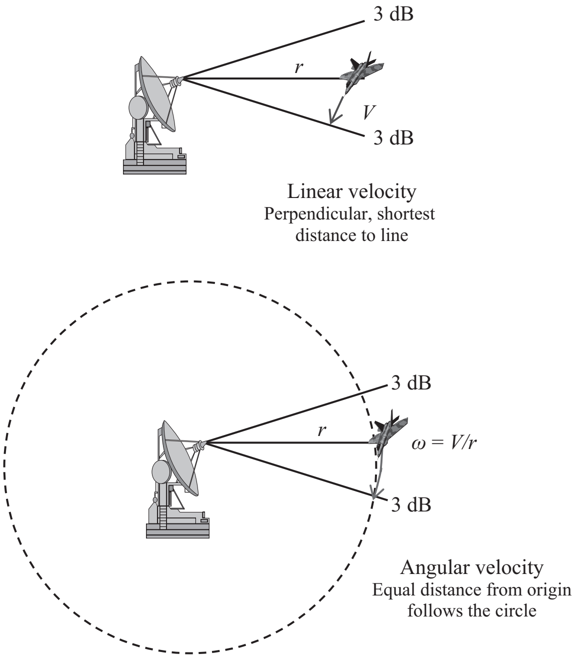


Figure 10.12 Linear and angular velocity

10.3.1 Linear and angular velocity

Linear velocity is the straight line velocity that a target will travel. It is the shortest distance between the boresite of the antenna beam and the 3 dB beamwidth, see Figure 10.12. This path is perpendicular to the 3 dB beamwidth and represents the shortest path that the target can take to fly out of the 3 dB beamwidth of the antenna from the boresite position.

Angular velocity is a curved line velocity around the platform with equal distance from the origin or platform antenna, see Figure 10.12. The angular velocity forms a circle around the origin. The angular velocity is equal to:

$$\omega = v/r$$

where ω is the angular velocity—radians/second, v is the linear velocity, and r is the range between the target and platform antenna or the radius of the circle.

For example, if a target is moving at 147 ft/s at a range of 672 ft, how fast does the tracker need to respond to keep the target in the main beam using a 3° beamwidth or ±1.5° (Figure 10.13).

Given: $v = 147$ ft/s, range $r = 625$ ft
Angular velocity $\omega = 147/625 = 0.235\ \mathrm{rad/s}$
Angular velocity $\omega = 0.235\ \mathrm{rad/s} \times 180/\pi = 13\ \mathrm{deg/s}$
Tracker speed = 1/2 beamwidth/angular velocity = 1/2 (3°)/13°/s = 115 ms

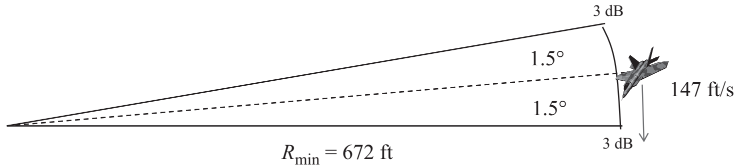


Figure 10.13 Angular velocity example

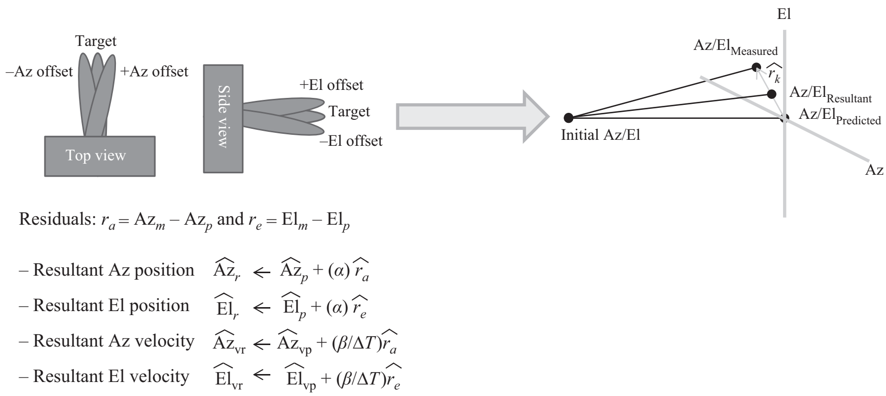


Figure 10.14 Two-dimensional sequential lobing

10.3.2 Sequential lobing

Sequential lobing is a method to provide the errors needed in track. This assumes that the lobing is in the beamwidth. This technique is very useful for AESAs that can switch rapidly to four positions on the mainlobe, +/−El and +/−Az. This lobing technique can be implemented for high-speed users with extreme accuracy and precision. The four-position sequential lobing +/−Az and +/−El is centered around the predicted, see Figure 10.14. This provides finer resolution than the sequential scan but is based on the same principle and technology. This technique points the antenna at four different locations and uses the RSS from each of the position to calculate which direction the antenna is pointed for boresite.

This four-beam lobing technique determines the Az- and El-measured position. It can be used for predicting tracking methods where it calculates difference between measured position and the predicted position as shown:

$$r_k = X_{k2\mathrm{Measured}} - X_{k2\mathrm{Predicted}}$$

This result updates the resultant Az and El positions for closed-loop tracking methods that will be discussed later.

It is also required to calculate/estimate processing delays to reduce lag between antenna command and measured RSS, which includes time stamping the RSS to ensure it matches the antenna position.

Sequential lobing needs to be analyzed for each application to ensure that the processing is fast enough to track the speed of the moving user platform. As the range becomes smaller, the required angular velocity increases in order to track inside the beamwidth. There are several parameters involved in determining the requirements for the tracker to stay in the main beam. These parameters are as follows:

- Angular velocity calculated by using the linear velocity and range.
- The beam positions required for the type of tracking, for sequential lobing, there are four beam positions.
- Number of samples needed for each beam position to receive an accurate measurement.
- The sample rate.
- Processing time or delay for all samples.
- Beamwidth of the mainlobe.
- Time to switch to the beam positions—not a consideration for AESAs.

For example:

Given:
Linear velocity = 735 ft/s, range = 1,000 ft
4 beam positions for sequential lobing
5 samples/beam position
Sample rate = 1 kH
Process delay = 5 ms
Beamwidth = 3°
Uses AESA for very high pointing speed (instantaneous, negligible)

Calculate angular velocity = 735/1,000 = 0.735 rad/s or 0.735 rad/s × 180/pi = 42°/s
Calculate total time delay = (#samples/beam position) × (#beam positions)/sample rate + process delay = $5 \times 4/1{,}000 + 5 \times 10^{-3}$ s = 25 ms
Calculate maximum angle error from boresite = angular velocity × total time delay = 42°/s × 25×10^{-3} = 1.05°
Required change of degrees to stay in mainlobe is beamwidth/2 = 3/2 = 1.5°
Calculate margin = required change − maximum angle error = 1.5—1.05 = 0.45°

If the margin is positive, then the target is tracked, if it is negative, then change is needed to track. This would require increasing the sample rate, reducing the number of samples, increasing the range, and/or widening the beamwidth.

10.3.3 CONical SCAN (CONSCAN)

A CONSCAN antenna creates a scan circle around boresite of the user. If the user platform is at boresite, then it is located in the center of the scan circle which means it has equal radius from the center of the scan circle to platform. Therefore, at boresite, the returned power will be equal at all angles, see Figure 10.15. If the user

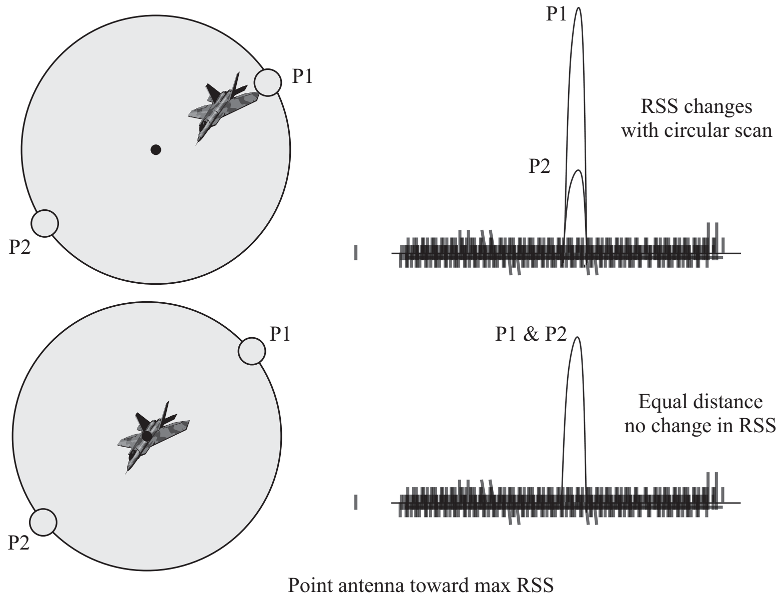


Figure 10.15 Conical scan basics

platform is off boresite, the power is greater toward the platform at a measured angle and is less at other angles of the scan, see Figure 10.15. This creates amplitude modulation (AM) of the received signal platform, and this AM is used to point the antenna directly at the user platform. The AM tells the antenna to move toward the highest amplitude, which uses both Az- and El-pointing vectors to move the antenna to the center of the scan until there is virtually no AM.

CONSCAN antennas are susceptible to interference and jamming, since they can receive external AM signals which results in errors for the CONSCAN solution. Atmospheric noise, blockage, jammers, other noise, and analogous propagation can be some of the sources of errors. Analogous propagation is the tunneling effect that received signal paths may encounter where amplitudes are much larger under certain conditions.

Jamming a conical scanner is relatively easy. The jammer sends out strong signals on the same frequency spoofing the desired return, for example, a series of random short bursts of signal that will appear to be a series of platforms in different locations within the beam. Improved or smart jammers time the signals to be the same as the rotational speed of the feed with a slight delay which results in a second strong peak within the beam, with nothing to distinguish the two receive levels. These types of jammers were used during World War II.

10.3.4 Monopulse

Monopulse is a single antenna divided into four quadrants or sections which uses measured angle and signal-level differences between the quadrants in order to point

the antenna toward the target. The antenna splits the beam into four sections and receives signals at slightly different directions, see Figure 10.16. The monopulse antenna determines which direction has a stronger signal which is the general direction of the platform relative to the boresight.

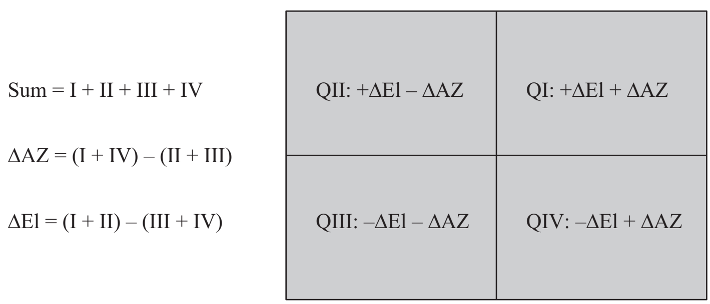
The monopulse antenna produces a sum of the received signal in four quadrants and two delta signals for Az and El, see Figure 10.16:

$$\begin{aligned}
\text{Sum} &= \mathrm{I} + \mathrm{II} + \mathrm{III} + \mathrm{IV} \\
\Delta\text{Az} &= (\mathrm{I} + \mathrm{IV}) - (\mathrm{II} + \mathrm{III}) \\
\Delta\text{El} &= (\mathrm{I} + \mathrm{II}) - (\mathrm{III} + \mathrm{IV})
\end{aligned}$$

The sum signal is for the transmit signal which corresponds with the antenna beam along center-line of the antenna. The sum signal is created by a feedhorn structure positioned to maximize signal at the center of the antenna beam.

The two delta signals are to receive, which produces the signals for El (up–down) and Az (left–right), see Figure 10.16. The pairs of quadrant beams are adjacent to the center-line of the sum antenna beam. This determines the RSS in the El and Az and which direction to point the monopulse antenna in order to ensure that the up and down amplitudes are the same, and the left and right amplitudes are the same for boresite. The differences in these delta channels produce the error and provide the information to move the antenna to boresite. The delta signals produce plus and minus results that determine the pointing direction of the antenna for tracking. When both delta signal equations are zero, the antenna is at boresite.

Monopulse antennas can also coast the last known value and velocity in order to predict the position of the user platform. It also can use techniques to avoid problems in decoding conventional conical scanning systems and prevents confusion caused by rapid changes in signal strength which helps to mitigate jammers.



Quadrants	Left	Right
Up	Q2: +ΔEl – ΔAZ	Q1: +ΔEl + ΔAZ
Down	Q3: –ΔEl – ΔAZ	Q4: –ΔEl + ΔAZ

Figure 10.16 Monopulse antenna

10.3.5 Alpha–beta tracker

Alpha–beta tracker uses the position state of the user platform and projects this position forward to predict its value at the next sampling time. The velocity V is constant for a small time interval ΔT between measurements. The variables α and β are selected according to the requirements. These values can also be empirically obtained for a given tracking system, but as a general rule, these values need to be positive and small:

$$0 < \alpha < 1; \; 0 < \beta < 2$$

There is a trade-off between convergence and error, the larger the numbers, the faster convergence time with greater error, the smaller these values are, the slower the convergence but provide a smaller error. In addition, these values may need to be updated according to the roll, pitch, and yaw information from the user platform.

As a general mode of operation, the corrections are small steps along an estimate of a parabolic gradient. The higher values provide faster convergence but higher residual error, the slower values provide slow convergence but with lower steady-state errors.

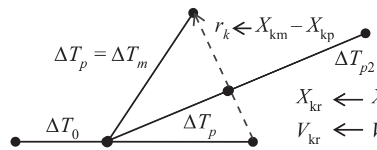
An example of an alpha–beta tracker is shown in Figure 10.17. The calculation steps are shown below:

1. Select α and β values with expected dynamics
2. Measure X_{k0}
3. Measure X_{k1}. Calculate initial velocity V_{k1}
4. Calculate predicted position X_{kp} and predicted velocity V_{kp}
5. Measure position X_{km} and calculate measured velocity V_{km}
6. Calculate the resultant vector r_k
7. Calculate resultant position X_{kr}
8. Calculate resultant velocity V_{kr}
9. Calculate next predicted position X_{kp2}
10. Calculate next predicted velocity V_{kp2}

Gamma can be added if the velocity is not sufficiently accurate and needs to consider acceleration, since the alpha–beta–gamma tracker uses position, velocity, and acceleration. In addition, Kalman filters can also be used for tracking and estimating positions. If the parameters are specified in azimuth and elevation, the same approach to the alpha–beta tracker can be used (Figure 10.18).

10.3.6 Integrating closed-loop tracker with open-loop navigational tracker

Open-loop navigational tracker is a location system using the signal data measured using GPS or other means to measure latitude, longitude, and altitude, from another external source. It is open-loop because it is a direct measurement of the data, and there is no feedback to determine position. The navigation tracker provides the measurements to calculate Az and El information for the antenna and is provided in platform's stabilized coordinate system. If the platform contains roll, pitch, or yaw,



Measure X_{k0}

Measure X_{k1}. Calculate initial velocity V_{k1}

Calculate predicted position X_{kp} and predicted velocity V_{kp}

Measure position X_{km} and calculate measured velocity V_{km}

Select α and β values with expected dynamics

Calculate the resultant vector r_k

Calculate resultant position X_{kr} using α

Calculate resultant velocity V_{kr} using β

Calculate next predicted position X_{kp2}

Calculate next predicted velocity V_{kp2}

X_{k0} = previous position
X_{k1} = initial position, V_{k1} = initial velocity
X_{kp} = predicted position, V_{kp} = predicted velocity
ΔT_p = delta time between initial and predicted positions
X_{km} = measured position, V_{km} = measured velocity
ΔT_m = delta time between initial and measured positions
X_{kr} = resultant position, V_{kr} = resultant velocity
r_k = difference between predicted and measured position vectors
Deviation from predicted due to noise and dynamic effects
α = position convergence multiplier $0 < \alpha < 1$
β = velocity convergence multiplier $0 < \beta < 1$
X_{kp2} = next predicted position, V_{kp2} = next predicted velocity
ΔT_{p2} = delta time between next predicted position and resultant position

Figure 10.17 Alpha–beta tracker for position prediction

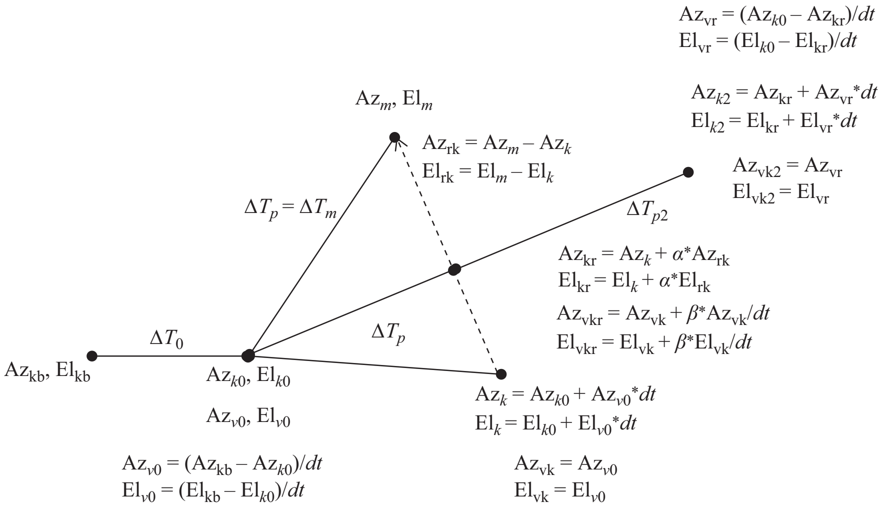


Figure 10.18 Alpha/beta using Az/El tracker

this information is needed to do a coordinate conversion in order to combine it with the closed-loop tracker, see Figure 10.19. This basically transforms the measurements of the open-loop tracker to the same coordinate system as the closed-loop tracker.

A closed-loop tracker is a location system using directional antennas which measure the RSS of the user and incorporates feedback to update and improve the tracking performance as well as predicting the location in the absence of a measured RSS. An example of a closed-loop tracker is the alpha–beta tracker discussed previously. The closed-loop tracker measures the pointing vector referenced to the rotated coordinates if roll, pitch, or yaw is present on the platform. Therefore, if these movements are present, the vector needs to be converted to the stabilized coordinate system before integration or if the open-loop tracker has already transformed the measurements to the nonstabilized baseline, then they can be integrated as is, see Figure 10.19.

Generally, the closed-loop tracker provides a more accurate position solution over the open-loop tracker and is used as the primary tracker if available. However, if the closed-loop tracker misses an RSS update measurement, it uses a coasting function to predict where the next position the user will be located. This is still generally a more accurate position; however, as more RSS measurements are lost, the open-loop tracker becomes more accurate.

Many systems simply switch over to the open-loop tracker when the closed-loop tracker has missed too many RSS measurements. However, a better solution would be to create an integrated solution with a weighting on the most accurate solution based on the number of measurements missed. This provides a smooth handoff between the closed-loop and the open-loop tracker and provides a more accurate estimate of the position.

Linear track integration solution

Coasting/Updates		αβ track					Nav track					Integrated Az	Integrated El
# Missed	# Good	W1	$\alpha\beta_{Az}$	$\alpha\beta_{El}$	$W1^*\alpha\beta_{Az}$	$W1^*\alpha\beta_{El}$	W2	Nav_{Az}	Nav_{El}	$W2^*Nav_{Az}$	$W2^*Nav_{El}$	$W1^*\alpha\beta_{Az} + W2^*Nav_{Az}$	$W1^*\alpha\beta_{El} + W2^*Nav_{El}$
0	5	1	3	4	3	4	0	5	7	0	0	**3**	**4**
1	4	0.8	3	4	2.4	3.2	0.2	5	7	1	1.4	**3.4**	**4.6**
2	3	0.6	3	4	1.8	2.4	0.4	5	7	2	2.8	**3.8**	**5.2**
3	2	0.4	3	4	1.2	1.6	0.6	5	7	3	4.2	**4.2**	**5.8**
4	1	0.2	3	4	0.6	0.8	0.8	5	7	4	5.6	**4.6**	**6.4**
5	0	0	3	4	0	0	1	5	7	5	7	**5**	**7**

Coast	Azimuth integration			El integration		
# of Coast	αβAz	NavAz	$W1^*\alpha\beta_{Az} + W2^*Nav_{Az}$	αβEl	NavEl	$W1^*\alpha\beta_{El} + W2^*Nav_{El}$
0	3	5	**3**	4	7	**4**
1	3	5	**3.4**	4	7	**4.6**
2	3	5	**3.8**	4	7	**5.2**
3	3	5	**4.2**	4	7	**5.8**
4	3	5	**4.6**	4	7	**6.4**
5	3	5	**5**	4	7	**7**

Logarithmic track integration solution for closed-loop track bias

Coasting/Updates		αβ track					Nav track					Integrated Az	Integrated El
# Missed	# Good	W1	$\alpha\beta_{Az}$	$\alpha\beta_{El}$	$W1^*\alpha\beta_{Az}$	$W1^*\alpha\beta_{El}$	W2	Nav_{Az}	Nav_{El}	$W2^*Nav_{Az}$	$W2^*Nav_{El}$	$W1^*\alpha\beta_{Az} + W2^*Nav_{Az}$	$W1^*\alpha\beta_{El} + W2^*Nav_{El}$
0	5	1	3	4	3	4	0	5	7	0	0	**3**	**4**
1	4	0.914	3	4	2.742	3.656	0.086	5	7	0.43	0.602	**3.172**	**4.258**
2	3	0.806	3	4	2.418	3.224	0.194	5	7	0.97	1.358	**3.388**	**4.582**
3	2	0.663	3	4	1.989	2.652	0.337	5	7	1.685	2.359	**3.674**	**5.011**
4	1	0.447	3	4	1.341	1.788	0.553	5	7	2.765	3.871	**4.106**	**5.659**
5	0	0	3	4	0	0	1	5	7	5	7	**5**	**7**

Coast	Azimuth integration			El integration		
# of Coast	αβAz	NavAz	$W1^*\alpha\beta_{Az} + W2^*Nav_{Az}$	αβEl	NavEl	$W1^*\alpha\beta_{El} + W2^*Nav_{El}$
0	3	5	**3**	4	7	**4**
1	3	5	**3.172**	4	7	**4.258**
2	3	5	**3.388**	4	7	**4.582**
3	3	5	**3.674**	4	7	**5.011**
4	3	5	**4.106**	4	7	**5.659**
5	3	5	**5**	4	7	**7**

Logarithmic track integration solution for Nav track bias

Coasting/Updates		αβ track					Nav track					Integrated Az	Integrated El
# Missed	# Good	W1	$\alpha\beta_{Az}$	$\alpha\beta_{El}$	$W1^*\alpha\beta_{Az}$	$W1^*\alpha\beta_{El}$	W2	Nav_{Az}	Nav_{El}	$W2^*Nav_{Az}$	$W2^*Nav_{El}$	$W1^*\alpha\beta_{Az} + W2^*Nav_{Az}$	$W1^*\alpha\beta_{El} + W2^*Nav_{El}$
0	5	1	3	4	3	4	0	5	7	0	0	**3**	**4**
1	4	0.553	3	4	1.659	2.212	0.447	5	7	2.235	3.129	**3.894**	**5.341**
2	3	0.337	3	4	1.011	1.348	0.663	5	7	3.315	4.641	**4.326**	**5.989**
3	2	0.194	3	4	0.582	0.776	0.806	5	7	4.03	5.642	**4.612**	**6.418**
4	1	0.086	3	4	0.258	0.344	0.914	5	7	4.57	6.398	**4.828**	**6.742**
5	0	0	3	4	0	0	1	5	7	5	7	**5**	**7**

Coast	Azimuth integration			El integration		
# of Coast	αβAz	NavAz	$W1^*\alpha\beta_{Az} + W2^*Nav_{Az}$	αβEl	NavEl	$W1^*\alpha\beta_{El} + W2^*Nav_{El}$
0	3	5	**3**	4	7	**4**
1	3	5	**3.894**	4	7	**5.341**
2	3	5	**4.326**	4	7	**5.989**
3	3	5	**4.612**	4	7	**6.418**
4	3	5	**4.828**	4	7	**6.742**
5	3	5	**5**	4	7	**7**

Figure 10.19 Coasting algorithm for no bias, closed-loop bias, and Nav track bias

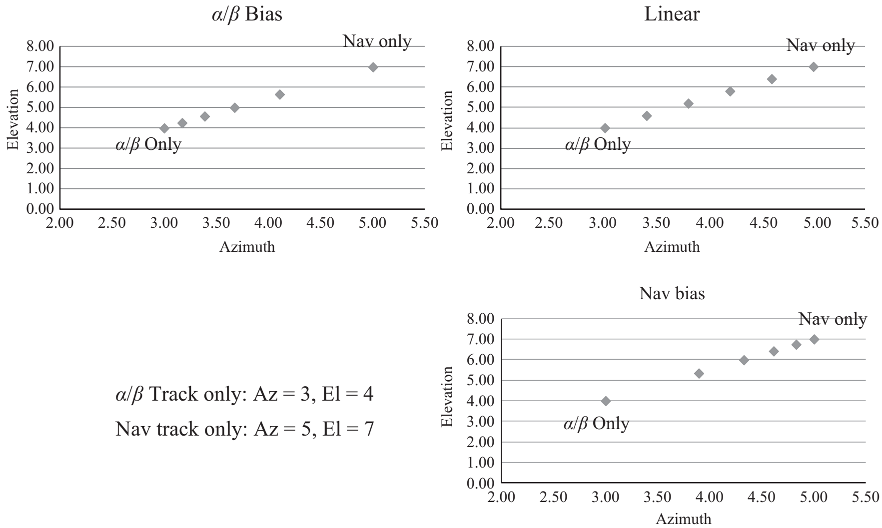


Figure 10.20 Track comparison using different bias for five allowable coasts for alpha–beta track and Nav track

An example of integrating the closed-loop alpha–beta tracker and the open-loop navigational tracker is shown in Figure 10.20. The nonbias example uses a simple linear integration. The close-loop bias uses logarithmically weighting that is, more toward the accuracy of the closed-loop system. The Nav track bias uses logarithmically weighting that is, more toward the accuracy of the open-loop system. Since the bias is selectable, the selection depends on the system used, accuracy of the sources, and dynamics of the users. The bias selected determines how the logarithmic weighting function is incorporated to provide the best accuracy during coasts, see Figure 10.20.

The number of coasts permitted in a system is variable and depends on several factors as discussed before. The number of coasts for the following example was five allowable coasts for the alpha–beta tracker and Navigation tracker integration. The biases for five coasts and the logarithmic weighting are shown in Figure 10.21. For the α/β bias, the solution remains closer to the α/β tracker solution, whereas the Nav bias solution shows that the solution moves over to the Nav solution faster, see Figure 10.21. To show the differences in the solutions of each of the biases, Figure 10.21 demonstrates the solutions of each of the biases with the changing number of coasts. After one coast, the solutions are close in value with the Nav bias being closest to the Nav-only solution. After two and three coasts, the solutions are spread out. After four coasts, the solutions are again close in value with the α/β bias closest to the α/β-only solution.

One variable that can determine which bias to select is the range of the user. For long ranges, use α/β bias. The Navigational tracker has the highest error at long range and also user dynamics are negligible at long range. For ranges greater than

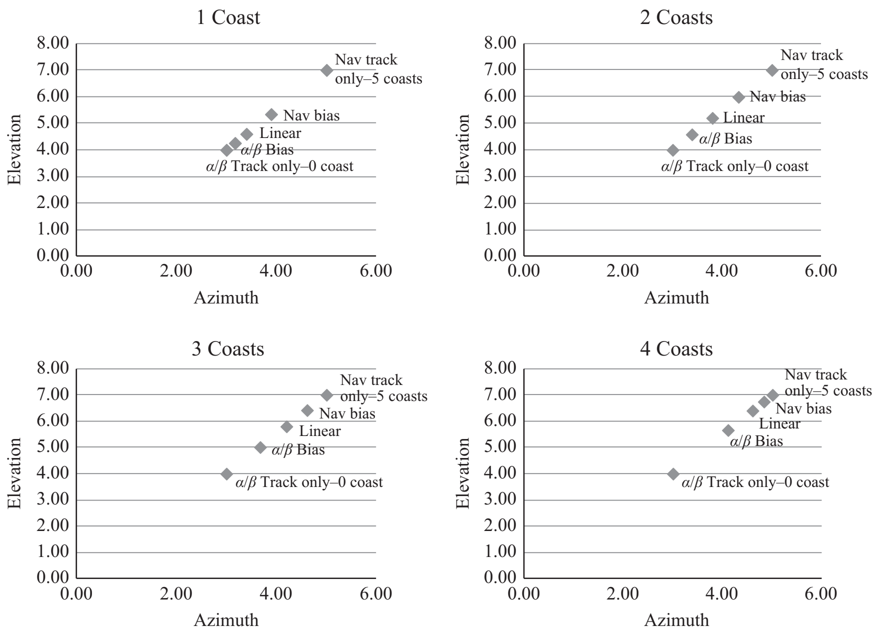


Figure 10.21 Coast example of five coasts for alpha–beta track and Nav track

20 nm, the α/β bias is used. For short ranges, the Nav bias is used, since the Nav error is small at short ranges and user dynamics are high; therefore, coasts for α/β tracker would produce larger errors.

The following steps are used to determine the position of the user:

1. Closed-loop tracker receives RSSI at a given Az/El position from the rotated antenna
2. Performs sequential lobing—calculates new Az/El pointing, sends command to rotated antenna
3. Calculates coasting Az/El position in absence of the RSSI, sends command to rotated antenna
4. Calculates Az/El antenna position on rotated coordinate system
5. Uses closed-loop Az/El and Nav data Az/El for integrated solution
6. Chooses either Nav bias or closed-loop bias for integration weighting
7. Sends integrated Az/El pointing command to the antenna

The weight values are calculated as follows for the different biases selected:

α/β bias:

α/β weight value is $= \log\{[(10-1)/(\text{total coasts})] \times (\text{total coasts} - \#\text{ of coasts}) + 1\}$

Nav weight value is $= 1 - \alpha/\beta$ weight

Nav bias:
Nav weight value is $= \log\{[(10-1)/(\text{total coasts})] \times (\#\text{ of coasts}) + 1\}$
α/β weight value is $= 1 -$ Nav weight

Linear bias:
α/β weight value is $= (1/\text{total coasts}) \times (\text{total coasts} - \#\text{ of coasts})$
Nav weight value is $= 1 - \alpha/\beta$ weight

10.3.7 Integrating external navigational track with internal navigational track

Integration of external Nav data with internal Nav data works basically the same way as the integrated solution for the closed-loop tracker and navigational data. The internal Nav data is the information sent from the user in the data link. It is generally more accurate than the external Nav data which uses an external source like a GPS unit. Therefore, the internal Nav data is used if it is available. If the internal Nav data is not available such as when the data link is lost or receives too many errors, then the system could switch to the external Nav data solution which would produce a less accurate solution. A better way to use both of these solutions is to coast the internal Nav solution for a given amount of coasts and then combine the weighted values of both the internal and external Nav solutions, see Figure 10.22. This again provides a smooth handoff from a more accurate to a less accurate positioning solution. The biases use weighting functions that are either weighted for the internal Nav solution, the external Nav solution, or linear with no weighting depending on the system and how reliable each of the solutions are, which also includes the dynamics of the user, see Figure 10.22. A comparison of the different bias results using five allowable coasts is shown in Figure 10.23. This shows that with the internal Nav bias, the results are closer to the internal Nav-only solution, and the external Nav bias shows the solution closer to the external Nav bias. The linear bias is evenly spaced for the transition between the internal Nav-only solution to the external Nav-only solution, see Figure 10.23.

The external Nav data generally uses the positioning information such as latitude, longitude, and altitude or pseudoranges from an external GPS. Many large platforms such as naval ships, aircraft, or military ground stations contain Link 16, UHF Omnidirectional link, or other data links to receive the external Nav-positioning information. The internal Nav data uses common data link where the Nav data is part of the data link.

Range and quality of the external Nav data can be used to determine which bias selection is used. For long range where the target dynamics are negligible, the internal Nav bias provides the most accurate positioning solution. A possible starting solution would be to use internal Nav bias for users that are >20 nm in range. The external Nav bias would be for users ≤ 20 nm where the Nav error is small at short ranges, and the target dynamics are high which would produce large errors due to amount of coasting of the internal Nav solution.

Linear track integration solution

Coasting/Updates		IN track					EN track					Integrated Az	Integrated El
# Missed	# Good	W1	IN_{Az}	$IN\beta_{El}$	$W1*IN_{Az}$	$W1*\alpha\beta_{El}$	W2	EN_{Az}	EN_{El}	$W2*EN_{Az}$	$W2*ENv_{El}$	$W1*IN_{Az} + W2*EN_{Az}$	$W1*IN_{El} + W2*EN_{El}$
0.00	5.00	1.00	3.00	4.00	3.00	4.00	0.00	5.00	7.00	0.00	0.00	**3.00**	**4.00**
1.00	4.00	0.80	3.00	4.00	2.40	3.20	0.20	5.00	7.00	1.00	1.40	**3.40**	**4.60**
2.00	3.00	0.60	3.00	4.00	1.80	2.40	0.40	5.00	7.00	2.00	2.80	**3.80**	**5.20**
3.00	2.00	0.40	3.00	4.00	1.20	1.60	0.60	5.00	7.00	3.00	4.20	**4.20**	**5.80**
4.00	1.00	0.20	3.00	4.00	0.60	0.80	0.80	5.00	7.00	4.00	5.60	**4.60**	**6.40**
5.00	0.00	0.00	3.00	4.00	0.00	0.00	1.00	5.00	7.00	5.00	7.00	**5.00**	**7.00**

Logarithmic track integration solution for internal Nav track bias

Coasting/Updates		IN track					Nav track					Integrated Az	Integrated El
# Missed	# Good	W1	IN_{Az}	$IN\beta_{El}$	$W1*IN_{Az}$	$W1*\alpha\beta_{El}$	W2	EN_{Az}	EN_{El}	$W2*EN_{Az}$	$W2*ENv_{El}$	$W1*IN_{Az} + W2*EN_{Az}$	$W1*IN_{El} + W2*EN_{El}$
0.00	5.00	1.00	3.00	4.00	3.00	4.00	0.00	5.00	7.00	0.00	0.00	**3.00**	**4.00**
1.00	4.00	0.91	3.00	4.00	2.74	3.66	0.09	5.00	7.00	0.43	0.60	**3.17**	**4.26**
2.00	3.00	0.81	3.00	4.00	2.42	3.22	0.19	5.00	7.00	0.97	1.36	**3.39**	**4.58**
3.00	2.00	0.66	3.00	4.00	1.99	2.65	0.34	5.00	7.00	1.69	2.36	**3.67**	**5.01**
4.00	1.00	0.45	3.00	4.00	1.34	1.79	0.55	5.00	7.00	2.77	3.87	**4.11**	**5.66**
5.00	0.00	0.00	3.00	4.00	0.00	0.00	1.00	5.00	7.00	5.00	7.00	**5.00**	**7.00**

Logarithmic track integration solution for external Nav track bias

Coasting/Updates		IN track					Nav track					Integrated Az	Integrated El
# Missed	# Good	W1	IN_{Az}	$IN\beta_{El}$	$W1*IN_{Az}$	$W1*\alpha\beta_{El}$	W2	EN_{Az}	EN_{El}	$W2*EN_{Az}$	$W2*ENv_{El}$	$W1*IN_{Az} + W2*EN_{Az}$	$W1*IN_{El} + W2*EN_{El}$
0.00	5.00	1.00	3.00	4.00	3.00	4.00	0.00	5.00	7.00	0.00	0.00	**3.00**	**4.00**
1.00	4.00	0.55	3.00	4.00	1.66	2.21	0.45	5.00	7.00	2.24	3.13	**3.89**	**5.34**
2.00	3.00	0.34	3.00	4.00	1.01	1.35	0.66	5.00	7.00	3.32	4.64	**4.33**	**5.99**
3.00	2.00	0.19	3.00	4.00	0.58	0.78	0.81	5.00	7.00	4.03	5.64	**4.61**	**6.42**
4.00	1.00	0.09	3.00	4.00	0.26	0.34	0.91	5.00	7.00	4.57	6.40	**4.83**	**6.74**
5.00	0.00	0.00	3.00	4.00	0.00	0.00	1.00	5.00	7.00	5.00	7.00	**5.00**	**7.00**

Figure 10.22 Coasting algorithm for all biases for integrating external and internal Nav track

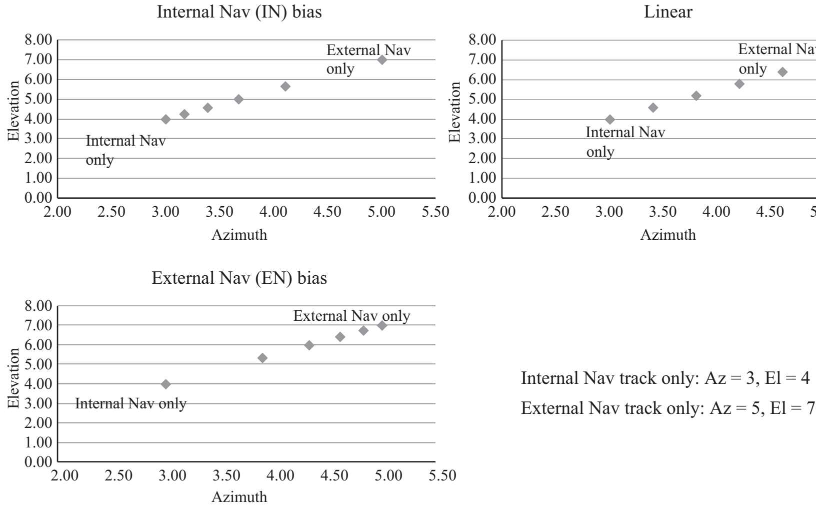


Figure 10.23 Track comparison using different bias for five allowable coasts for external and internal Nav track

Another factor that would dictate which of the biases would be used is the quality of external Nav information. This depends on the external source of the data received. If the positioning data has poor quality of the positioning data, then internal Nav bias would be selected. However, if the quality of the external Nav bias is better, then external Nav bias would be used.

10.4 Summary

Volume search, acquisition, and track are important technologies which need to be incorporated in order to find and track moving user platforms using directional antennas. These are important to any directional wireless data link, and several methods have been discussed with providing trade-offs of the optimal solutions. Random cell search method provides the fastest response to find a user in a volume space. The inside-out and outside-in methods are optimal using previous positions of the users and also to prevent losing a user during the search.

Integrated positioning solutions are required to provide the optimal handoff and coasting between users. This provides the best accuracy and also provides smooth handoff between coasting a very accurate solution with a measured less accurate method.

10.5 Problems

1. List 3 processes that are used to find and monitor a moving platform.
2. What discovery method uses the shortest time to discover a moving platform?

3. What discovery method is the best to prevent losing the moving platform?
4. What is the best type of volume search to use if the previous user position was known?
5. What two antenna techniques can be used to scan a great volume regardless of the type of scan?
6. How much change in E_b/N_o occur when the antenna beam is spoiled to achieve 4 times the beamwidth which decreases the search time by 1/4? Can the signal still be detected at the same range? Why?
7. What antenna method is used to acquire a user?
8. What antenna method is used to track a user?
9. Describe how a conical scan CONSCAN works?
10. What does it mean when the CONSCAN AM is zero?
11. Describe how a sequential scanning works? How does it find a target in two dimensions?
12. Describe how sequential lobing works for tracking?
13. What is a simple tracking method that only uses position and velocity?
14. What is the difference between an open-loop tracker and a closed-loop tracker?
15. What method is used to provide a smooth handoff between the closed-loop tracker and the navigation open-loop tracker?
16. What type of bias or weighting would generally be used for long range? For short range?

Further reading

Skolnik, Merrill. *Introduction to Radar Systems*, 3rd ed. New York City, NY: McGraw-Hill, 2002.

Chapter 11
Broadband communications and networking

Broadband refers to technology that distributes high-speed data, voice, and video. It is used for broadband communications, networking, and Internet access and distribution and provides a means of connecting multiple communication devices. Several generations of wireless communication products—designated as 1G (first generation) through 5G (fifth generation)—have evolved over the years and will continue to advance in the future.

Broadband is also used in the home to connect to the outside world without having to run new wires. In this application, information is brought to the home in various ways including power lines, phone lines, wireless radio frequencies (RFs), fiber-optic cable, coaxial cable, and satellite links. This information generally comes into the home at a single location, so it is necessary to have a way to distribute it throughout. Along with the distribution of information, networking plays an important role in the connection and interaction of different devices in the home.

The military is investigating in several networking techniques such as the Joint Tactical Radio System (JTRS) and Link 16 to allow multiple users in a battle scenario for communications, command, control, and weapon systems.

11.1 Mobile users

Wireless communications have progressed from simple analog cellular telephones to high-speed voice, data, music, and video using digital communications techniques. 1G devices provide wireless analog voice service using advanced mobile phone service (AMPS) but no data services or Internet connection.

2G devices use digital communications with both voice and data capability. The data rate for 2G is from 9.6 to 14.4 kbps. Several techniques are used to provide 2G communications, including code division multiple access (CDMA), time division multiple access (TDMA), global system for mobile communications (GSM), and personal digital cellular (PDC) for one-way data transmission. This type of connection does not have always-on data connection, which is important for Internet user applications.

3G is an enhanced version of the 2G digital communications with an increased data rate of 114 kbps to 2 Mbps; always-on data for Internet applications; broadband services such as video, music, and multimedia; superior voice quality; and

Table 11.1 Generations of mobile radios

Generation	Data rate	Modulation access	Additional features
1G	Analog	AMPS	Voice, no data or internet services
2G	9.6–14.4 kbps	GSM, CMDA, TDMA, PDC	Voice, data, no always-on for internet
3G	114 kbps–2 Mbps	CDMA (W-CDMA), CDMA-2000, TD-SCDMA	Always-on for internet, video, music, multimedia, voice quality, enhanced roaming
4G	1 Gbps	SDRs, OFDM, MIMO, FDMA, turbo codes, cognitive radios	Adaptation for interoperability and real-time modulation changes
5G	>1 Gbps	Higher frequency bands: 28, 37, 39 GHz	Increased spectral efficiency, better coverage, enhanced efficiency, reduced latency, more users with higher data rates

enhanced roaming. The three basic forms of 3G are wideband CDMA (W-CDMA), CDMA-2000, and a time division scheme called time division synchronous CDMA (TD-SCDMA).

4G provides a means of adapting the communication link so that it is interoperable with multiple communication techniques and higher data rates up to 1 Gbps. This is accomplished using software-defined radios (SDRs) for real-time modulation changes and cognitive radio (CR) techniques that sense the environment and adjust the modulation and other parameters according to the signals present.

5G is being developed with the modulation techniques to be defined. The advantages that are projected are increased spectral efficiency, better coverage, enhanced efficiency, reduced latency, and more users operating with higher data rates. In addition, frequency band allocations include 28, 37, and 39 GHz.

A summary of the five generations is shown in Table 11.1.

11.1.1 Personal communications services

The cellular telephone used analog techniques for many years before the creation of personal communications services (PCS). This technique uses digital communication modulation to send digital data wirelessly. To enable multiple users in the same frequency band, CDMA, TDMA, and frequency division multiple access (FDMA)—or a combination of any of these multiple access schemes—were implemented (Figure 11.1). The Federal Communications Commission (FCC) allowed companies to develop their own multiple access solution and chose not to impose a standard for the development of PCS. This caused an issue with interoperability. To combat this issue, two main solutions emerged. The first, CDMA, was developed and adopted early on by several companies as a standard for sending

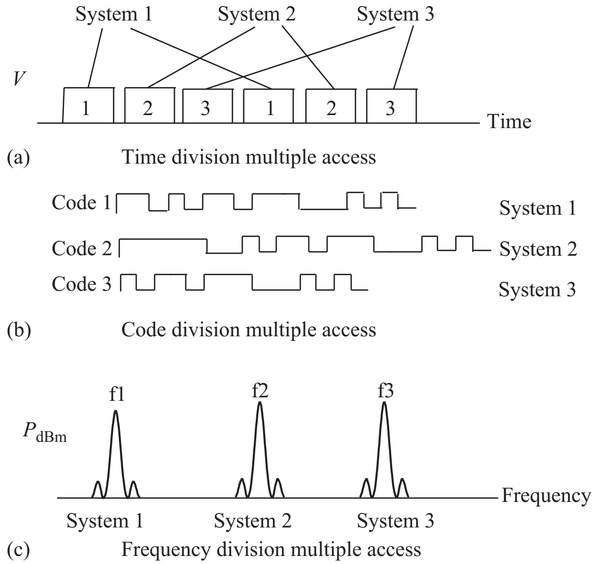


Figure 11.1 Multiple access methods for PCS communications

and receiving digital communications. The second, GSM, was patterned after the standard in many European countries. Most PCS hardware accommodates one of these two approaches. Many PCS telephones are capable of using more than one technique. Also, other companies competing for the interoperability "standard" used TDMA and a combination of CDMA/TDMA.

Some systems have both PCS and cellular modes to allow the end user a choice and to provide additional coverage and versatility. Since the onset of PCS hardware, the cellular industry also has developed digital communication techniques to enhance its performance and to compete with PCS. Some systems incorporate PCS, digital, and analog cellular all in one handset, but the trend now is to eliminate the analog and old cellular devices, which are obsolete, and to focus on the PCS technology and beyond.

GSM. The GSM standard originated in Europe and has been used for years, which has allowed for many improvements. The standards in Europe include GSM-900 and DCS-1800. In addition, the Digital European Cordless Telecommunications (DECT) in Europe set a popular standard that provides digital wireless communications and the DECT wireless telephone.

The United States followed the standard set in Europe by using GSM technology, but at a slightly different frequency. The standard for GSM in the United States was called PS-1900, which operates at 1,900 MHz band.

Modulation for GSM. Gaussian minimum shift keying (see Chapter 2) is used to modulate GSM system because of its reduced sidelobes and out-of-band transmissions, making it a spectrally efficient waveform with its ability to send data at a high rate for digital communications. The bandwidth specified to send voice or data

over the air is 200 kHz. This allows multiple users to be on the band at the same time with minimal interference. During this process, known as FDMA, the users occupy a portion of the band that is free from interference. GSM also incorporates frequency-hopping (FH) techniques, controlled by the base to further prevent multiple users from interfering with each other.

The specification for GSM makes it possible to send control bits to set power output, which also helps to prevent interference from other users since they should use only the minimum power required for a reliable and quality connection. The power control helps the near–far problem, which occurs in a handset that is close to the base. If the handset is allowed to transmit at full power, then it will jam most of the other handsets. If power control is used, then the handset's power is reduced to the necessary power needed. This prevents jamming and saturation of the base and allows for handsets that are farther away from the base unit to turn up the power and reach the base. The optimum system would be to have the power level close to the same level regardless of the distance from the base station. The specification for GSM also allows adjustment of the duty cycle.

The data rate for GSM is approximately 277 kbps, with a 13-kbps vocoder. Two types of GSM are specified: Reflex 25, which uses two frequency channels; and Reflex 50, which uses four frequencies for twice the data speeds. The rate of change from one frequency to another is 2.4 kHz. Digital signal processors (DSPs) reduce the distortion in the overall system and detect the information reliably.

11.1.2 Cellular telephone

Cellular telephones have been around for several decades. The first ones used an analog frequency modulation system that operated in the 800-MHz band under AMPS. With the arrival of PCS digital telephones, cellular needed to transition to digital modulation techniques to compete in the marketplace. Cellular had the advantage of an already established and well-proven infrastructure and allocated antenna sites. The PCS industry had to start from scratch. Cellular companies developed a digital modulation system and combined it with analog technology to create what they called the dual mode AMPS. They also implemented TDMA IS-54/IS-136 and CDMA techniques as digital cellular operation.

11.1.3 Industrial, scientific, and medical bands

The FCC designated frequency bands to be used solely by industrial, scientific, and medical (ISM) organizations (Table 11.2). These ISM bands were chosen for the PCS community as secondary users with strict requirements to not interfere with the existing primary users of the bands. This included the use of spread spectrum techniques to minimize interference.

With the development of PCS telephones, the FCC opened up the ISM bands for digital PCS communications with requirements that need to be met to use these bands for wireless telephony communications. These requirements also apply to other applications such as broadband communications and home networking, which

Table 11.2 Industrial, scientific, and medical bands used for communications

ISM bands	
Freq range	**Center freq**
13.533–13.567 MHz	13.560 MHz
26.957–27.283 MHz	27.120 MHz
40.66–40.70 MHz	40.68 MHz
902–928 MHz	915 MHz
2.4–2.5 GHz	2.45 GHz
5.725–5.875 GHz	5.8 GHz
24–24.25 GHz	24.125 GHz

are now using the ISM bands for RF communications. The following ISM bands are now popular for use with RF solutions:

902–928 MHz
2.4–2.5 GHz
5.725–5.875 GHz

Also, ISM higher frequency bands 24 to 24.5 GHz are currently used for residential use in fixed wireless systems.

11.2 Types of distribution methods for the home

Three types of distribution methods or mediums are used to distribute high-speed voice, data, and video throughout the home: transmission over power lines, transmission over phone lines, and transmission through the air using RF communications.

11.2.1 Power-line communications

A power-line communication (PLC) system modulates voice, data, and video signals and sends them over the existing alternating current (AC) power lines in the home. This method uses the infrastructure of the existing power lines inside the home and provides very broad coverage without adding additional wires. Also, since the hardware requires coupling to the power lines but no antennas or high-frequency components, this method is relatively low cost.

The power line medium itself is fairly noisy. In addition, the different lengths of wiring and different terminations lead to several impedance mismatches, which cause peaks and nulls in the amplitude response of the spectrum (Figure 11.2). Thus, most systems use sophisticated digital transmission design methods to mitigate the problems with medium and frequency agility to prevent operation in amplitude nulls.

Orthogonal frequency division multiplexing. One of the more common techniques used for PLC is orthogonal frequency division multiplexing (OFDM; see Chapter 2). This method uses multiple frequencies, each of which is orthogonal, to

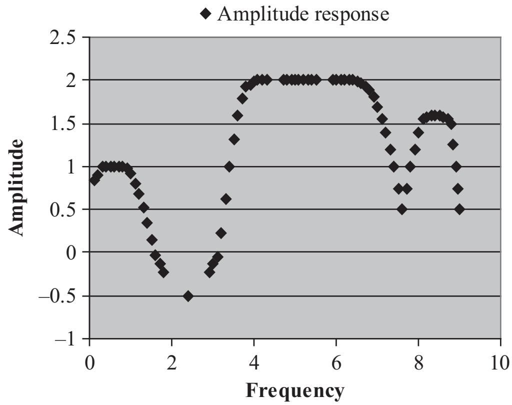


Figure 11.2 Spectrum of the amplitude response of the power line

send the data. This allows independent detection of all the frequency channels used, with minimum interference between channels. OFDM is used in conjunction with digital modulation schemes such as binary phase-shift keying (BPSK), quadrature phase-shift keying (QPSK), and quadrature amplitude modulation (QAM). This allows higher throughput of the data utilizing multiple channels in parallel and allows for a very robust system to drop frequency channels that are being jammed or are unusable due to the amplitude spectral response.

11.2.2 Home phoneline networking alliance

The home phoneline networking alliance (PNA) transmits signals over the existing telephone line infrastructure. Therefore, wherever there is a telephone line present, voice and data can be transmitted without running additional wires. Home PNA has two standards: 1.0 and 2.0. The 1.0 standard is capable of 1 Mbps. The home PNA is used for shared Internet access, printer and file sharing, and network gaming. The 2.0 standard increased the speed to 11 Mbps and must be backward-compatible with the 1 Mbps 1.0 standard. The home PNA is essentially an Ethernet network that uses the telephone lines for infrastructure. Home PNA developers now provide USB and Ethernet adapters to the home PNA network and have increased the data rates greater than 100 Mbps.

The data and voice are coupled to the telephone line and transmitted to other devices via the telephone lines. This system is ideal if the existing home has multiple telephone lines. In many homes, however, the telephone line infrastructure does not provide a means to distribute signals into every room in the house. This is a major limitation to home PNA. The power line and RF methods provide a much better solution for full coverage throughout the home.

11.2.3 Radio frequency communications

RF is a method for distributing voice, data, and video throughout the home. It provides the required bandwidth to extend high-definition television (HDTV) and

Table 11.3 Home networking systems

Home networking system	Implementation	Advantages	Disadvantages
Power line	Apply modulation on A/C power lines Utilizes OFDM	Power lines in all rooms and garage No new wires Low cost	Power line noise Variable impedances and multiple terminations
Phone line	Apply modulation on the telephone lines	Phone lines existing	Not all rooms have telephone lines Limited coverage
RF	RF modulation GMSK, ISM bands 802.11, Bluetooth, WiMAX, LTE, others	Good coverage Adaptable to external wireless devices	Multipath, RF interference

other broadband signals and offers connectivity to external wireless devices. However, RF can be degraded by multipath and interference from other RF signals.

RF technologies exhibit problems with multipath, blockage, and band saturation from other users, causing nulls in the amplitude spectrum and distorted signals. Spread spectrum schemes and antenna diversity can help to mitigate these issues. Networking standards including the IEEE 802.11 standard and Bluetooth are established and verified. Table 11.3 summarizes the different home networking methods. RF distribution is the most commonly used and has the best performance.

11.2.4 Wi-Fi (IEEE 802.11)

The IEEE 802.11 standard specifies both the physical (PHY) layer and the medium access control (MAC) layer and has been adopted as a standard for home RF networking. The PHY layer can use both RF and infrared (IR). RF uses spread spectrum techniques, including either FH or direct sequence, and operates in the 2.4-GHz ISM band. The 802.11 standard specifies initial data rates of either 1 or 2 Mbps and has increase the throughput well over 100 Mbps. Further developments include operation in the 5-GHz band with much higher data rates.

The MAC layer uses the carrier sense multiple access with collision avoidance (CSMA/CA) protocol. This permits the network to have multiple users and reduces the chance of transmittal at the same time. Each node checks to see if the channel is being used and if not it transmits data. If the channel is busy, it randomly selects the amount of time that it waits before transmitting again. Since there are multiple time selections, the probability is very low that two nodes trying to enter the network will select the same random time slot, which further reduces the chance of collisions.

To ensure that a connection is made between two nodes, the transmitting node sends out a ready-to-send (RTS) packet, and the receiving node, upon successful

reception of the RTS packet, sends a response of clear-to-send (CTS), which lets the transmitting node know that it is communicating with that particular node and not with another one. Then, the transmitting node sends the data, and the receiving node acknowledges reception by returning an acknowledge (ACK) to the transmitter, which is verified by a cyclic redundancy check (CRC). Wi-Fi networks are short-range, high-bandwidth networks primarily developed for data.

Their quality of service (QoS) is similar to fixed Ethernet, where packets can receive different priorities based on their tags. For example, Voice over Internet Protocol (VoIP) traffic may be given priority over web browsing.

Wi-Fi runs on the MAC's CSMA/CA protocol, which is connectionless and contention based. The 802.11 specification defines peer-to-peer (P2P) and ad hoc networks, where an end user communicates to users or servers on another local area network (LAN) using its access point (AP) or base station. Wi-Fi uses the unlicensed spectrum, generally 2.4 GHz, using direct sequence spread spectrum, multicarrier OFDM to provide access to a network. Subscriber stations that wish to pass data through a wireless AP are competing for the AP's attention on a random interrupt basis, which causes distant subscriber stations from the AP to be repeatedly disturbed by closer stations, greatly reducing their throughput.

Wi-Fi is supported by most personal computer operating systems, many game consoles, laptops, smartphones, printers, and other peripherals. If a Wi-Fi-enabled device such as a PC or mobile phone is in range, it connects to the Internet. In addition, both fixed and mobile computers can network to each other and to the Internet. The interconnected APs, called hotspots, are used to set up the mesh networks. Routers that incorporate DSL or cable modems and a Wi-Fi AP give Internet access to all devices connected (wirelessly or by cable) to them. They can also be connected in an ad hoc mode for client-to-client connections without a router. Wi-Fi enables wireless voice applications—Voice over Wireless LAN (VoWLAN) or Wireless VoIP (WVoIP)—and provides a secure computer networking gateway, firewall, Dynamic Host Configuration Protocol server, and intrusion detection system. Any standard Wi-Fi device will work anywhere in the world. The current version of Wi-Fi Protected Access encryption (WPA2) is not easily defeated.

QoS is more suitable for latency-sensitive applications (voice and video), and power-saving mechanisms (WMM Power Save) improve battery operation.

Wi-Fi's distance is limited to typically 32 m (120 ft) indoors and 95 m (300 ft) outdoors. The 2.4-GHz band provides slightly better range than 5 GHz. In addition, directional antennas can increase the distance to several kilometers. Zigbee and Bluetooth are in the range of 10 m but continue to increase.

Wireless network security is a concern for Wi-Fi users. Attackers gain access by monitoring Domain Name Service (DNS) requests and responding with a spoofed answer before the queried server can reply. The security methods for Wi-Fi include the following:

1. Suppress AP's Service Set Identifier (SSID) broadcast. SSID is broadcast in the clear in response to client SSID query. However, since it is also transmitted for other signals, this is not a valid security method.

2. Allow only computers with known MAC addresses to join the network. Unfortunately, MAC addresses are easily spoofed.
3. Use Wired Equivalent Privacy (WEP) encryption standard. This can be easily broken.
4. Use Wired Protected Access (WPA) with TKIP. However, more secure is no longer recommended.
5. WPA2 encryption standards are still considered secure and are used presently.
6. Future encryption standards will be forthcoming due to the sophisticated attacks on security.

Wi-Fi APs default with no encryption and provide open access to other users. Security needs to be configured by the user via graphical user interface (GUI). Other security methods include virtual private networks (VPNs) or a secure Web page.

Unless it is configured differently, Wi-Fi uses 2.4 GHz APs that default to the same channel on startup. This causes interference if there are multiple users. In addition, high-density areas and other devices at 2.4 GHz can cause interference with many Wi-Fi APs. An alternative is to use the 5-GHz band, but the interference will increase as more and more users operate in that band.

Wi-Fi uses an AP that connects a group of wireless devices to an adjacent wired LAN for Internet connection or networking. The AP relays the data between wireless devices and a single wired device using an Ethernet hub or switch. In this way, wireless devices can communicate with other wired devices. Wireless adapters are used to permit devices to connect to a wireless network. The connection to the devices is accomplished using external or internal interconnects such as a PCI, USB, ExpressCard, Cardbus, and PC card. Wireless routers integrate WPA, Ethernet switch, and internal router firmware applications to provide Internet Protocol (IP) routing, network address translation, and DNS forwarding through an integrated WAN interface. Wired and wireless Ethernet LAN devices connect to a single WAN device such as a cable or DSL modem. This allows all three devices (mainly the AP and router) to be configured through one central utility or integrated web server that provides web pages to wired and wireless LAN clients and to WAN clients.

11.2.5 Bluetooth (IEEE 802.15)

Bluetooth is a standard that operates at 2.4 GHz in the ISM band (2.4–2.4835 GHz). The band allows 79 different channels with a channel spacing of 1 MHz. It starts with 2.402 GHz, which allows for a guard band of 2 MHz on the lower band edge, and ends with 2.480 GHz, which allows for a guard band of 3.5 MHz on the upper band edge. These band allocations and guard bands are the US standards. Other countries may operate at slightly different frequencies and with different guard bands.

The following three types of classes deal mainly with power output and range:

- Class 1: +20 dBm
- Class 2: +4 dBm
- Class 3: 0 dBm

The higher the power output provides a greater range for the system. However, both Class 1 and 2 require power control to allow only the required power to be used for operation. Class 1 operation is required to reduce the power to less than 4 dBm maximum when the higher power output is not required. The power control uses step sizes from 2 dB minimum to 8 dB maximum. The standard method of power control uses a received signal strength indication (RSSI). Depending on the RSSI in the receiver, this information is sent back to the transmitter to adjust the transmitted power output.

The modulation scheme specified in the Bluetooth standard uses differential Gaussian frequency-shift keying (DGFSK) with a time bandwidth (BT) of 0.5 and a modulation index between 0.28 and 0.35, with "+1" being a positive frequency deviation and "−1" being a negative frequency deviation. The frequency deviation is greater than 115 kHz. The symbol rate is 1 Msym/s. Full duplex operation is accomplished using packets and time division multiplexing, and for this application, it is referred to as time division duplexing (TDD).

FH is used for every packet, with a frequency hop rate of 1,600 hops/s, providing one hop for every packet. The packets can cover from one to five time slots, depending on the packet size. Each time slot is 625 μs long, which corresponds to the hop rate. However, since the frequency hop rate is dependent on the packet size, if the packet is longer than the slot size, then the hop rate will decrease according to the packet size. For example, if a packet covers five time slots, then the hop rate would be 1/5 of the 1,600 hops/s, or 320 hops/s, which equals 3.125 ms dwell time, which equals 5×625 μs.

The standard packet is made up of three basic sections: access code, header, and payload (Figure 11.3). The access code is 72 bits long and is used for synchronization, direct current (DC) offset compensation, and identification. The detection method used is a sliding correlator. As the code slides in time against the received signal, when the code matches up to the received signal the correlation of the two produce a peak, and when that peak is greater than a set threshold, it is used for a trigger in the receiver timing.

Bluetooth forms networks that are either point-to-point (PTP) or point-to-multipoint (PMP). PMP is called a piconet and consists of a master unit and up to seven slave units, which can be in multiple piconets, called scatternets, using TDM. However, the piconets cannot be time or frequency synchronized.

There are three types of access codes: channel access code, device access code, and inquiry access code. The channel access code identifies the piconet. The device

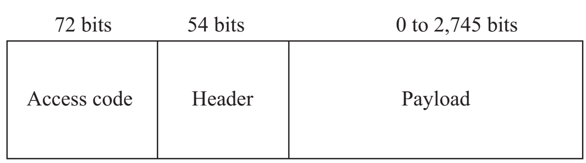


Figure 11.3 Basic structure of a standard packet defined in the Bluetooth specification

access code is used for applications such as paging. The inquiry access code is for general inquiry access code and checks to see what Bluetooth devices are within range. The dedicated inquiry access code is used for designated Bluetooth units in a given range. The header is 54 bits long and determines what type of data (e.g., link control) is being sent. A wide variety of information types are specified in the Bluetooth specification. The payload is the actual data or information that is sent. Depending on the amount of data, this can range from 0 to 2,745 bits in length.

Three error correction schemes are used with Bluetooth: one-third-rate forward error correction (FEC), two-third-rate FEC, and automatic repeat request (ARQ). Error detection is accomplished using CRCs. The standard also specifies a scrambler or a data whitener to prevent long constant data streams (all "1"s for a period of time or all "0"s for a length of time), which cause problems in carrier and tracking loops.

The capacity of the Bluetooth standard can support an asynchronous data channel, asynchronous connectionless link, three PTP simultaneous synchronous voice channels, synchronous connection-oriented links, or an asynchronous data channel and a synchronous voice channel. The synchronous voice channel is 64 kbps for quality voice in both directions. The asymmetrical speeds for Bluetooth are up to 723.2 and 57.6 kbps on the return channel or 433.9 kbps if both links are symmetric.

The Bluetooth system is required to have a minimum dynamic range of 90 dB. The receiver for Bluetooth is required to have a minimum sensitivity of −70 dBm and needs to operate with signals up to +20 dBm.

The Bluetooth (802.15), Wi-Fi (802.11), and Worldwide Interoperability for Microwave Access (WiMAX) (802.16) and LTE specifications are summarized in Table 11.4. The initial Bluetooth specification was noted for its short range and was used mainly for device control applications. With the increase in power output, up to +20 dBm, its range has increased so that it can now be used for multiple applications. However, Bluetooth was not designed for high-speed data rates and will eventually need to be redesigned or replaced by devices that are focused on high-speed data rates and broadband communications.

11.2.6 WiMAX (IEEE 802.16)

WiMAX is an infrastructure that uses PMP links to provide mobile Internet users a connection to the Internet Service Provider (ISP), for example, a wireless laptop computer. Using a PTP connection to the Internet, users gain access via cell phones to a fixed location. WiMAX provides P2P and ad hoc networks, which give end users the ability to communicate with other users or servers on another LAN using its AP or base station.

WiMAX is the standard for broadband wireless access with the following options:

802:16d: fixed WiMAX
802:16e: mobile WiMAX
802:16m: WiMAX II for 4G

Table 11.4 RF commercial standards

RF system	Performance	Waveform/ Security	Modulation	Data rate	Other
IEEE 802.11 Wi-Fi 802.11b/g—mobile	RF or infrared (IR) ubiquitous 100 m fixed range APs provide Internet-access—"Hot Spots"	ISM band 2.4 GHz protected access encryption (WPA2) attacks via DNS	Spread spectrum direct sequence frequency hop	1 or 2 Mbps future 11 Mbps CCK	CSMA/CA protocol. Compete for access point AP random interrupt basis near/far problem
Bluetooth IEEE 802.15	FEC, ARQ, CRC Sensitivity: −70 dBm DR = 90 dB Saturation: +20 dBm	ISM band 2.4 GHz 79 channels 1 MHz spacing PTP, PMP, TDM FEC, Scrambler	Class 1–3: +20 dBm +4 dBm, 3: 0 dBm DGFSK BT= 0.5 Mod Index 0.28–0.35 FH1,600 h/s, 1 h/pktet Full duplex—TDM/TDD	Symbol rate −1 Msps Synch—64 kbps Asynch—723.2 kbps	Full duplex TDD ACL, SCO PTP PMP
WiMAX—Worldwide interoperability for microwave access 802.16d—fixed 802.16e—mobile 802.16m—WiMAX II for 4G	ISP to end user Voice, 4G, TDD, FDD, TRANSEC, FIPS, DIA CAP, JITC 4G mobile broadband	Multiple bands PMP & PTP BWs: 1.4–20 MHz 200 active users in every 5-MHz cell	BPSK, QPSK 16 & 64 QAM TDD, TD-SCDMA GSM/GPRS, CDMA, W-CDMA (5 MHz) UMTS 3GPP2 networks—cdmaOne & CDMA2000	2–15 Mbps (original) Downlink 100 Mbit/s, Uplink 50 Mbit/s	OFDMA MIMO Time slots assigned Resolves near/far
LTE – long-term evolution Voice, 4G, GSM/GPRS, CDMA, W-CDMA (5 MHz) UMTS	TDD, FDD, Security: 3% overhead, TRANSEC, FIPS, DIA CAP, JITC 4G mobile broadband	Multiple bands Flexible bandwidths—1.4 to 20 MHz 200 active users in every 5-MHz cell	QPSK, 16 & 64 QAM TDD, TD-SCDMA GSM/GPRS, CDMA, W-CDMA (5 MHz) UMTS 3GPP2 networks—cdmaOne & CDMA2000	Downlink 100 Mbit/s, Uplink 50 Mbit/s	MIMO, OFDM SC-FDMA Turbo codes interleaving

A WiMAX forum was established in June 2001 to ensure conformity and interoperability.

The bandwidth and range of WiMAX can provide the following functions:

1. Connection to Wi-Fi hotspots to the Internet.
2. The "Last mile" connection to replace or enhance cable and DSL.
3. Both data (for computers) and telecommunications (wireless telephone) services.
4. Internet connectivity backup to fixed services such as cable and DSL.
5. Broadband access to the Internet.

WiMAX is part of the Radio Communications Sector (RCS) of International Telecommunication Union (ITU-R) or IMT-2000. It provides international interoperability for mobile communications. Current mobile communications that use WiMAX include handsets (e.g., cellular smartphones, Android handset from Google, 3G EV-DO devices, Qualcomm's IPR); PC peripherals (e.g., PC cards or USB dongles); and laptops and electronics devices (e.g., game terminals, MP3 players).

Using WiMAX, cable companies can gain access to wireless networks such as the mobile virtual network operator, which has the ability to move between the Clearwire and Sprint 3G networks. It replaces GSM and CDMA and the wireless backhaul for 2G, 3G, 4G, and 5G networks.

The North America Backhaul for urban cellular consists of a copper T1 line, and for remote places, a satellite is used. In most other regions, urban and rural backhaul is provided by microwave links such as local multipoint distribution service (LMDS) or multichannel multipoint distribution service (MMDS). Since copper wire is not adequate for broadband communications, the microwave backhaul provides 34 Mbps and 1 Gbps speeds with latencies in the order of 1 ms.

The WiMAX MAC/data link layer consists of a base station that assigns the subscriber an access slot upon entry that can be increased or decreased depending on demand or priorities. The scheduling algorithm allows users to compete for the time slot, but once they are in a permanent access, slot is allocated to them. This provides stability under overload conditions and is much more bandwidth efficient than other methods. The base station controls the QoS by balancing time slot assignments with application needs of subscriber stations. The connection is based on specific scheduling algorithms.

The WiMAX physical layer is certified, which gives vendors interoperability. The frequency range is 10 to 66 GHz, with some standard frequencies being 2.3, 2.5, and 3.5 GHz (Table 11.4). The backhaul generally uses frequencies in the 5-GHz unlicensed band. In addition, the analog TV bands (700 MHz) are being used for WiMAX applications. Details of the WiMAX data link are shown in Table 11.4.

The available bandwidth is shared between users, so with more users, the individual performance can deteriorate. Most users can expect around 2–3 Mbps. Multiple-in, multiple-out (MIMO) techniques are used for increased coverage.

WiMAX allows for self-installation, power consumption, frequency reuse, and bandwidth efficiency. Fast Fourier transforms (FFTs) are used to provide scaling to

the channel bandwidth for constant carrier spacing of 1.25, 5, 10, or 20 MHz. WiMAX provides higher spectrum efficiency in wide channels and cost reduction in narrow channels using the following multiple techniques:

- Intelligent dynamic frequency selection (I-DFS) advanced antenna diversity
- Hybrid automatic repeat-request (HARQ)
- Adaptive antenna systems (AAS) and MIMO technology
- Turbo coding and low-density parity check (LDPC)
- Downlink subchannelization for coverage versus capacity
- Extra QoS class for VoIP applications
- Stationary and mobile (4G, 5G)
- Dynamic burst algorithm, which adjusts burst profile for signal-to-noise ratio (SNR) at the same time
- Modulation agility, which has more bits per OFDM/scalable orthogonal frequency division multiple access (SOFDMA) symbol for high SNR, less bits for low SNR, more robust burst profile
- Integration with an IP-based network, which gives a defined connection with an IP-based core network at the ISP and a base station that works with other types of architectures
- Packet switched mobile networks

WiMAX maintains a flexible architecture to allow remote and mobile stations of varying scale and functionality and to provide base stations of varying size (e.g., femto, pico, and mini as well as macro).

11.2.7 LTE

LTE is a next generation of communication and networking. LTE is currently based on 4G radio technology, including TDD and frequency division duplex (FDD). Security for LTE requires approximately 3% overhead and contains the following: Transmission Security (TRANSEC), *Federal Information Processing Standards* (FIPS), Department of Defense Information Assurance Certification and Accreditation Process (DIACAP), and Joint Interoperability Test Command (JITC). This technology currently uses 4G mobile broadband and all-IP flat architecture with data rates (peak) of 100 Mbps for the downlink and 50 Mbps for the uplink. The latency is less than 10 ms for round-trip time and less than 5 ms latency for small IP packets. If MIMO is used, the peak download rates are 326.4 Mbps for 4×4 antennas and 172.8 Mbps for 2×2 antenna for every 20 MHz of bandwidth. The peak upload rate is 86.4 Mbps for every 20 MHz. LTE provides flexible bandwidths of 1.4 to 20 MHz with low operating expenditures (OPEX). TDD in LTE is aligned with TD-SCDMA as well, which provides coexistence and seamless connection with GSM/General Packet Radio Service, CDMA, W-CDMA, Universal Mobile Telecommunications System UMTS, and 3rd Generation Partnership Project 3GPP2 networks including cdmaOne and CDMA2000 (IS-95 CDMA).

LTE supports multicast broadcast single frequency network (MBSFN) for mobile TV. It uses an evolved packet system and evolved UMTS Terrestrial Radio

Access Network (UTRAN) (evolved E-UTRAN) on the access side and evolved packet core and system architecture evolution (SAE) on the core side.

This technology uses five different terminal classes, from voice-centric to high-end terminal, and 200 active users in every 5 MHz cell (200 active data clients). It also supports W-CDMA at 5 MHz and has an optimal cell size of 5 km with 30 and 100 km cell sizes being supported.

The 3GPP, which uses 3G technology, proposed the Transmission Control Protocol/Internet Protocol (TCP/IP) with studies into all IP networks (AIPN). The higher level protocols such as LTE SAE are flat network architectures. They are efficient supports of mass-market usage of any IP-based service and are the evolution of the existing GSM/W-CDMA core network. The Release 8 E-UTRA for UMTS operators is geared to be used over any IP network, WiMAX, Wi-Fi, and other wired networks.

3GPP uses orthogonal FDMA (OFDMA) for the downlink (tower to handset) and single carrier FDMA (SC-FDMA) for the uplink. It also incorporates MIMO with up to four antennas per station. Turbo coding is used along with a contention-free quadratic permutation polynomial (QPP) turbo code internal interleaver. OFDM is used for downlink.

The time domain data structure is as follows:

- Radio frame = 10 ms
- 10 subframes of 1 ms each
- Subframe contains two slots
- Each slot = 0.5 ms; subcarrier spacing = 15 kHz

Twelve of these subcarriers together (per slot) are called a resource block:

- One resource block = 180 kHz
- 6 resource blocks fit in a carrier of 1.4 MHz
- 100 resource blocks fit in a carrier of 20 MHz

The Physical Downlink Shared Channel (PDSCH) includes all the transmitted data. It uses various types of modulations including QPSK, 16-state QAM (16-QAM), and 64-QAM, depending on the required data throughput and the available SNR. The Physical Multicast Channel is the broadcast transmission using a single frequency network (SFN). The Physical Broadcast Channel sends the most important system information within the cell.

Single-user MIMO is used to increase the user's data throughput. Multi-user MIMO is used for increasing the cell throughput.

The uplink is a precoded OFDM called an SC-FDMA. This is to compensate for high peak-to-average power ratio (PAPR) for normal OFDM requiring high linearity. SC-FDMA groups resource blocks to reduce linearity requirements. There are two physical channels in the uplink signal: Physical Random Access Channel (PRACH) for initial access and when the UE is not uplink synchronized; and Physical Uplink Shared Channel (PUSCH) for all data using QPSK, 16-QAM, and 64-QAM. The MIMO provides spatial division multiple access and increases

the data rate depending on the number of antennas that are used. This in turn ensures that more than one mobile unit may reuse the same resources.

The following has been demonstrated with the results shown:

- HDTV streaming >30 Mbps, video supervision, and mobile IP-based handover between LTE and High-Speed Downlink Packet Access (HSDPA)
- Bit rates up to 144 Mbps
- Data rates of 200 Mbps with <100 mW
- HD video streaming with peak rates of 96 Mbps downlink, 86 Mbps uplink, and VoIP
- Data rates Mbps traveling 110 km/h
- M700 mobile platform peak data rates 100 Mbps downlink and 50 Mbps uplink
- 100 Mbps uplink transfer speeds

Most carriers currently supporting GSM or HSPA networks will eventually upgrade to LTE network. Companies are upgrading to LTE as their 4G technology but will provide an interim solution using HSUPA and HSPA+. Upgrade paths for CDMA networks will also migrate to LTE.

11.3 Local multipoint distribution service

LMDS is a PMP distribution service that allocates communication signals with a relatively short RF range to multiple end users. In this multipoint system, the base station or hub transmits signals in a PMP method that is basically a broadcast mode. The return path from the subscriber to the base station or hub is accomplished via a PTP link.

The LMDS system is basically an end-to-end microwave radio link that enables VoIP, local and long-distance telephony services, high-speed Internet access and data transmissions, and video conferencing and TV. It uses digital wireless transmission systems operating in the 28-GHz range for systems in the United States and 24–40 GHz for overseas systems, depending on the area and country. LMDS can be employed in either an asymmetric or symmetric configuration. The asymmetric system has a data rate that is different depending on the direction of the transmission, for example, the downstream has a higher data rate than the upstream, as is the case with asynchronous digital subscriber loop. The symmetric configuration has the same data rate regardless of which direction the data are sent. The FCC has assigned frequency as blocks to the end user. Block A users are assigned 1,150 MHz of bandwidth in different frequency bands:

Frequency	Bandwidth
27.5–28.35 GHz	850 MHz
29.1–29.25 GHz	150 MHz
31.075–31.225 GHz	150 MHz
Total bandwidth	1,150 MHz

Block B users are left with the remainder of the bandwidth, which equals 150 MHz total. This total bandwidth allocation of 1,300 MHz for LMDS represents the largest FCC bandwidth allocation of the wireless data communications systems including PCS, cellular, direct broadcast satellite, MMDS, digital audio radio services, wireless communications service, and interactive and video data and over two times the bandwidth of AM/FM radio, very high frequency and ultra-high frequency TV, and cellular telephone combined. Most of the above systems are allocated a fraction of the LMDS allocation, for example, the ISM band for PCS at the 900-MHz band is only 26 MHz. The standard intermediate frequency (IF) for typical LMDS systems ranges from 950 to 2,050 MHz.

Many LMDS systems use the concepts of bandwidth-on-demand and shared resources by integrating their systems with asynchronous transfer mode (ATM). IP transport methodology is also used to send data and VoIP. LMDS is used to span the "last mile" to the user's facilities. These systems transmit very high data rates up to 500 Mbps each way. However, the distance is limited to about 2 to 4 mi. One of the reasons for this short range can be attributed to the rain fade at these frequencies of up to 30 dB of attenuation in the link analysis. Using the bandwidth allocated to LMDS, these systems can provide up to 85 Mbps to the residential users and up to 155 Mbps to the commercial users. A typical LMDS can provide up to 155 Mbps downstream and a return link of 1.544 Mbps. These high-speed data rates are compatible with the T1 speed requirements up to and including OC-3 connections. Using just the first part of the Block A bandwidth of 850 MHz and using QPSK modulation, this system can provide 100 simultaneous T1 lines. With the additional bandwidth and higher order modulation schemes, this capacity can be increased.

All communication systems involved with providing multiple user access require a method of multiplexing these users so that they can all access the same system. The basic multiplexing schemes used are CDM/CDMA, TDM/TDMA, and FDM/FDMA. The difference in the multiplexing methods and the multiple access methods is that the multiplexing method assigns the code, time slot, or frequency to a user regardless of whether the user is using the system, whereas with multiple access schemes, the user is assigned a code, time slot, or frequency as needed when the user accesses the system. This assignment is only for that period of use, and then that code, time slot, or frequency is released and made available for the next user after the original user exits the system.

Most LMDS systems today use either FDM/FDMA or TDM/TDMA to provide multiple access for a given system. FDMA works well for applications where users are connected on a more continuous basis, and TDMA is better for when users are more periodic. If two users require the link on a continuous basis, TDMA would not be suitable because they would have to share the time allocation, but with FDMA, the time is always allocated to each of the users and the access is separated in frequency. If the two users are periodic in their usage and do not require continuous access to the system, then TDMA might be the method to use because it takes advantage of when the subscriber is not using the services to increase either the number of users on the system or the overall data rate of the connection. These

factors and others are considered when deciding on which type of multiplexing is optimum. Many times both of these methods are used in a system to provide better access, more users, and higher speeds for each user with minimal interference between them.

FDMA may be the best solution to large customers since they are using the Internet on a fairly continuous basis. However, for small end users, the requirement is more periodic. They require high usage for downstream applications while they are downloading large files, but generally the usage and speed requirements are lower during upstream use. Therefore, to provide an optimum use system, when the demand is high, and when the demand is low, a TDM/TDMA scheme is generally used.

Orthogonal methods and spatial antenna separation also make it possible to have multiple users. Some examples of orthogonal techniques include OFDM, orthogonal phase systems like QPSK, and orthogonal antenna polarizations, using vertical and horizontal polarizations for different users. Satellite communications currently use this technology.

Spatial antenna separation, or space sectoring, is another multiple access scheme that allows multiple subscribers to use the same communication system without jamming each other by having directional antennas pointed to different space segments of the intended users. The number of users and spatial separation depends on the beam width and the amount of isolation between beams.

To optimize a communications link, various combinations of orthogonal techniques, multiple access schemes, and spatial antenna separation are used to provide the optimum solution and the maximum amount of end users per system.

LMDS uses different types of digital modulation schemes depending on complexity and bandwidth efficiency. Some of the more popular modulation schemes include BPSK, QPSK, O-QPSK, 8PSK, using both standard and differential on any of the modulation types, and various QAM systems including 16-QAM and 64-QAM. The higher the order of modulation scheme, the higher the data rates for a given bandwidth. However, these systems are generally more complex, require higher signal levels (or have a reduce range), and are costlier to implement.

Modulation techniques provide a higher data rate for bandwidth efficiency, and multiple access techniques allow multiple users to access the system. The combination of these methods is used to maximize the capacity of a given system by providing the maximum users per base station or site. The higher capacity that can be achieved for a given base station provides more coverage or less base stations per given coverage. Therefore, the modulation and multiple access schemes are carefully designed for each system installed.

Since LMDS systems have a relatively short RF range, the base stations or hubs are spaced a few kilometers apart and are linked together to provide service up to several thousand end users. The main reason for reduced coverage in LMDS systems is atmospheric conditions, mainly rain. Their advantages over other wireless systems is that they are line of sight (LOS), that their antennas are fixed at the sight, that they are usually mounted at a high elevation (rooftops), and that often the

antennas are directional so multipath and blockage are generally not a problem after the antennas are mounted and operational. However, the atmosphere is constantly changing, which affects the range of the system. Antenna position and mounting are important factors along with the type of modulation used to provide the range required for coverage and capacity.

Four elements make up an LMDS system: the customer premises equipment (CPE), which is the end user of the system; the base station or hub, which services multiple end users; the fiber-based infrastructure, which is the wired connection to the central office (CO); and the network operations center (NOC), which provides the networking and can operate with or without a CO.

Examples of CPEs include LANs, telephones, faxes, Internet, video, television, and set-top boxes (STB). The interfaces to these devices are digital signal, DS-0, DS-1 structured and unstructured T1/E1, T3/E3, DS-3, OC-1, OC-3/STS-3 fiber optics, ATM and video communications, plain old telephone service (POTS), frame relay, Ethernet 10BaseT/100BaseT, and others. Both scalable and nonscalable network interface units (NIUs) at the CPE are used to go between the incoming LMDS signals and the devices being used at the CPE. The scalable NIU is used in large business and commercial uses. It is a flexible system that can be configured for the application, and it is chassis based. Therefore, the same chassis can be used for many types of systems and applications.

The main elements of the NIU are the modem and the data processor that supports the different type of external connections or interfaces. The nonscalable unit is for small and medium-sized users and is for a fixed application and interfaces. This unit is designed specifically for certain types of interfaces: T1/E1, T3/E3, POTS, 10BaseT, video, frame relay, and ATM.

The network-node equipment (NNE) connects the wireline functions to the LMDS wireless link and contains the processing, multiplexing/demultiplexing, compression/decompression, modem, error detection and correction, and ATM.

The base station or hub provides the interface between fiber infrastructure and wireless infrastructure. The elements of a standard base station are antenna system and microwave equipment for both transmitting and receiving, downconverter/upconverter, modulators/demodulators, and interface to the fiber. The base station antennas are mounted on rooftops and other high places to prevent blockage from structures since the LMDS frequencies are LOS.

Local switching in the base station can permit communications between users operating in the same network without going through the wired infrastructure if a NOC is provided. The fiber-based infrastructure is the wired connections mainly to the CO. This provides the connections from the base stations or NOC to the local CO.

The NOC consists of the network management system equipment consisting of optical network (SONET) optical carrier OC-12 and OC-3, DS-3 links, CO equipment, ATM and IP switching systems, and interconnections with the Internet and public switched telephone networks. Some implementation schemes do not use a NOC, and the information from the base station is connected through fiber to ATM switches or CO equipment at the CO so that all communications must go

through the CO. By using a NOC, if users on the same network desire to communication, they can do so directly through the NOC bypassing the CO.

The current applications for LMDS are for both the suburban and rural communities throughout the world where it becomes difficult and impractical to run copper wire. In large cities, the buildings and roads are already established, and there is no practical way of running copper wire except to tear up roads and alter buildings, which becomes very disruptive, costly, and impractical. This is the case for many large cities in the United States and also many worldwide cities. Also, in many of the smaller cities, this can pose a major problem, requiring the same disruptions and demolitions. Since LMDS is a wireless solution, it can be installed on rooftops of buildings without having to run wires or tear up roads and will have a minimum impact to the community while simultaneously providing high-speed communications through its extended bandwidth.

For rural communities where the population is scarce, it becomes impractical and costly to provide services to the few that live in rural communities. In addition, they are spread out so that the infrastructure of a wired solution becomes more costly running miles of copper wire and often times is not made available to these customers. The LMDS offers a solution by providing a wireless link to these remote communities at a lower cost and less time to install. Throughout the world in many remote, underdeveloped countries, there does not currently exist a wired infrastructure for use with telephone, Internet, and video applications. LMDS can provide a solution for these countries by setting up a wireless infrastructure that is less costly and requires less time to install. Often in these countries, it takes years after a request is made to provide a hard-wired line to these end users, and sometimes the request is never fulfilled. LMDS not only can provide an immediate solution to the problem but also can extend the coverage that the wired lines provide. This enhances the coverage of the hard-wired fiber and coax lines to these remote places.

Presently, most of the applications are corporate because the cost to implement them residentially is too high. However, over the next several years, they will become more viable for home use.

11.4 MMDS

MMDS is a wireless service that operates from 2.2 to 2.4 GHz. It has a range of approximately 30 mi, which is LOS at these frequencies. It was originally designed to provide one-way service for bringing cable TV to subscribers in remote areas or in locations that are difficult to install cable. Other systems use bandwidths of 200 MHz in the band just above 2.5 GHz and also in the Ka-band at 24 GHz. The power output allowed is up to 30 W, and OFDM is used to enhance the number of users and increase the speeds. This provides up to 10 Mbps during peak use and can provide speeds up to 37.5 Mbps to a single user.

MMDS has the capacity to support up to 33 analog channels and more than 100 digital channels of cable television. Although MMDS was mainly designed for this

cable TV service where wiring was impractical and costly to run, in 1998, FCC passed rules to allow MMDS to provide data and Internet services to subscribers. MMDS is used as a short-range inexpensive solution.

The IEEE 802.16 working group for broadband wireless access networks sets the requirements for both the MMDS and LMDS and other wireless "last mile" type technologies. Also, OFDM and VOFDM technology has been proposed for use in the MMDS system solution. The lower frequency bands, 2.5 and 5.0 GHz, which fall into the MMDS bands, are more appealing at the present time due to cost, availability, and quickness to install. LMDS may provide a long-term future solution to the "last mile" connection, but it will take more time to get the infrastructure in place for this LOS technology. The International Telecommunications Union (ITU) and the European Telecommunications Standards Institute (ETSI) are also involved in setting standards for LMDS and MMDS.

11.5 Universal mobile telecommunications system

UMTS is a 3G mobile networking that provides wide-area wireless cellular voice telephone, video calls, and broadband wireless data with data rates up to 14.4 Mbps on the downlink and 5.8 Mbps on the uplink using W-CDMA and GSM. It supports up to 21 Mbps data transfer rates with HSDPA and up to 42 Mbps HSDPA+. UMTS, using 4G technology, operates 100 Mbps downlink and 50 Mbps uplink using OFDM. In addition, UMTS combines W-CDMA, TD-CDMA, or TD-SCDMA, GSM's mobile application part core, and the GSM family of speech codecs. The modulation is 16-QAM operation at a frequency of 1,710–1,755 and 2,110–2,155 MHz. Users access the 3G broadband services using a cellular router, Personal Computer Memory Card International Association, or USB card.

A 3G phone can be used as a gateway or router to provide connection of Bluetooth-capable laptops to the Internet.

11.6 4G

4G provides wireless communications, networking, and connection to the Internet for voice, data, multimedia messaging service, video chat, mobile TV, HDTV, and digital video broadcasting. 4G is a spectrally efficient system that gives high network capacity and high throughput data rates: 100 Mbps for on the move and 1 Gbps for fixed location. It also provides a smooth handoff across heterogeneous networks, seamless connectivity, and global roaming across multiple networks. 4G contains high QoS for next-generation multimedia support of real-time audio, high-speed data, HDTV video content, and mobile TV. It has been designed to be interoperable with existing wireless standards, which include an all-IP, packet switched network. 4G systems dynamically share and utilize network resources to meet the minimal requirements. They employ many techniques to provide QoS such as OFDM, MIMO, turbo codes, and adaptive radio interfaces as well as fixed

relay networks and the cooperative relaying concept of multimode protocol. The packet-based (all-IP) structure creates low latency data transmission.

4G technology is incorporated into LTE and the higher speed version of WiMAX (WiMAX IEEE 802.16m) for 3GPP, which uses OFDMA, SC-FDMA, and MC-CDMA.

Interleaved FDMA (IFDMA) is being considered for the uplink on the 4G system since OFDMA contributes more to the PAPR. IFDMA provides less power fluctuation and thus avoids amplifier issues. It needs less complexity for equalization at the receiver, which is an advantage with MIMO and spatial multiplexing. It is capable of high-level modulations such as 64-QAM for LTE.

IPv6 support is essential when a large number of wireless-enabled devices are present because it increases the number of IP addresses, which in turn provides better multicast, security, and route optimization capabilities.

Smart or intelligent antennas are used for high rate, high reliability, and long range communications. MIMOs provide spatial multiplexing for bandwidth conservation, power efficiency, better reception in fading channels, and increased data rates.

11.7 Mobile broadband wireless access IEEE 802.20

The Mobile Broadband Wireless Access standard covers mobile units moving at 75 to 220 mph (120 to 350 km/h). High-speed dynamic modulation and similar scalable OFDMA capabilities are utilized to handle these speeds and provide fast handoff, FEC, and cell edge enhancements. The mobile range is approximately 18 mi (30 km) with an extended range of 34 mi (55 km). The data rate for WiMAX2 and LTE going forward is 100 Mbps for mobile applications and 1 Gbps for fixed applications.

11.8 MIMO communications

MIMO is used to provide spatial multiplexing (SMX) using two antennas at both ends. One antenna transmits one data bit, and another antenna transmits another bit simultaneously; this makes up one symbol. The receiver contains two antennas to separate the signals and receives the two bits and the same time. The receiver then multiplexes these bit streams into a data stream that provides a data rate at twice the original data rate transmitted out of each antenna with no increase in bandwidth. Therefore, the data rate is twice as fast compared with using STC with only one receive antenna.

MIMO can be used to improve the link and robustness. Using MIMO for a more robust link, a single data stream is replicated and transmitted over multiple antennas. The redundant data streams are each encoded using a mathematical algorithm known as space time block codes. With such coding, each transmitted signal is orthogonal to the rest, which reduces self-interference and improves the capability of the receiver to distinguish between the multiple signals. With the

multiple transmissions of the coded data stream, there is increased opportunity for the receiver to identify a strong signal that is less adversely affected by the physical path. MIMO is fundamentally used to enhance system coverage (Matrix A).

MIMO can also be used to increase data throughput. The signal to be transmitted is split into multiple data streams and each data stream is transmitted from a different base station transmit antenna operating in the same time-frequency resource allocated for the receiver. In the presence of a multipath environment, the multiple signals will arrive at the receiver antenna array with sufficiently different spatial signatures, allowing the receiver to readily discern the multiple data streams. Spatial multiplexing provides a very capable means for increasing the channel capacity (Matrix B).

This technique can be used to increase the data rate with adding the number of antennas. For example, WiMAX uses four antennas at both ends with either a four times increase in data rate, improved signal quality and twice the data rate, or additional improvement in signal quality with the same data, which reduces error rates caused by blockage, multipath, or jamming.

MIMO or MISO can be configured for multiple users that are spatially separated. This allows users to be on the same frequency with the base using multiple directional antennas, each of them pointed in the direction of each of the users. In addition, synchronizing the users can provide further improvement using TDMA and the multiple antennas.

Active electronically steered arrays (AESAs) can be used to provide AAS for multiple beams in a single antenna or aperture. In addition, beamforming can be used to shape the beam for the best reception. Another way to improve performance is via cyclic delay diversity, in which multiple signals are delayed before transmission. At the receiver, these signals are combined using the specified delays for a more robust received signal. The closer the signal can get toward a flat channel at a certain power level, the higher the throughput.

11.9 MISO applications for cellular networks

MISO used in cellular network implementations contains multiple antennas at the base station and a single antenna on the mobile device, which is often referred to as space time code (STC). This improves reception, range, data rate, and spectral efficiency. Autonegotiation is used dynamically between each individual base station and mobile station to adjust specific parameters such as transmission rates. In addition, multiple mobile stations can be supported with different MISO capabilities as needed, which maximize the sector throughput by leveraging the different capabilities of a diverse set of vendor mobile stations.

This technique uses different data bit constellations (phase/amplitude) that are transmitted on two different antennas during the same symbol. The conjugate or inverse of the same two constellations is transferred again on the same antennas during the next symbol. The data rate with STC remains the same with the received

signal being more robust using this transmission redundancy. Similar performance can be achieved using two receive antennas and one transmitter antenna, which is generally referred to as antenna diversity and relies on the fact that while one of the antennas is blocked or in a multipath null, the other antenna receives a good signal.

11.10 Quality of service

QoS is a measure of the guaranteed level of performance regardless of capacity. It uses different priorities for different applications, users, or data rates. It monitors the level of performance and dynamically controls the scheduling priorities in the network to guarantee the performance by this scheduling, or it can use the reserve capacity if needed. The required bit rate, delay, jitter, packet dropping probability, and bit error rate are part of the QoS. A possible alternative to QoS is to provide margin to accommodate expected peak use.

QoS has requirements for many parameters such as service response time, signal loss, signal-to-noise ratio, cross-talk, echo, interrupts, frequency response, and loudness levels. The grade of service is used in the telephony world and relates to capacity and coverage such as maximum blocking probability and outage probability.

QoS can be the cumulative effect on subscriber satisfaction. Generally, four types of service bits and three precedence bits are provided in each message that are redefined as DiffServ Code Points (DSCPs) for the modern Internet. For packet-switched networks, QoS is affected by human factors such as stability of service, availability of service, delays, and user information and also technical factors such as reliability, scalability, effectiveness, and maintainability.

Some of the problems that occur in transmissions include the following:

1. Dropped packets. Can prevent communications and possible retransmission of the packets causing delays.
2. Delay. Excessive delay due to long queues or a less direct route can cause problems with real-time communications such as VoIP.
3. Jitter. Variation in delay of each packet can affect the quality of streaming audio or video.
4. Out-of-order delivery. Different delays of the packets can cause the packets to arrive in a different order causing problems in video and VoIP.
5. Error. Distorted or corrupted packets causing errors in the data, needs error detection or ARQ.

Many applications require QoS, including streaming multimedia, IPTV, IP telephony or VOIP, and video teleconferencing (VTC).

There are basically two types of QoS: inelastic, which requires a minimum bandwidth and maximum latency, and elastic, which is versatile and can adjust to the available bandwidth. Bulk file transfer applications that rely on TCP are generally elastic. A service-level agreement is used between the customer and the provider to guarantee performance, throughput, and latency by prioritizing traffic.

Resources are reserved at each step on the network for the call as it is set up; this is known as the Resource Reservation Protocol (RSVP). Commercial VoIP services are competitive with traditional telephone service in terms of call quality without QoS mechanisms. This requires margin above the guaranteed performance to replace QoS, and it depends on the number of users and the demands.

Integrated services (IntServ), which reserve the network resources using RSVP to request and reserve resources through a network, are not realistic because of the increase in users and the demand for more bandwidth. Differentiated services (DiffServ) use packets marked according to the type of service needed. This service uses routers for supporting multiple queues for packets awaiting transmission. The packets requiring low jitter (VoIP or VTC) are given priority over packets in the other queues. Typically, some bandwidth is allocated by default to the network control packets. Additional bandwidth management mechanisms include traffic shaping or rate limiting.

The TCP rate control artificially adjusts the TCP window size and controls the rate of ACKs being returned to the sender. Congestion avoidance is used to lessen the possibility of port queue buffer tail drops, which also lowers the likelihood of TCP global synchronization. QoS deals with these issues concerning the Internet. The Internet relies on congestion-avoidance protocols, as built into TCP, to reduce traffic load under conditions that would otherwise lead to Internet meltdown. Both VoIP and IPTV necessitate large constant bitrates and low latency, which cannot use TCP. QoS can enforce traffic shaping that can prevent it from becoming overloaded. This is a critical part of the Internet's ability to handle a mix of real-time and nonreal-time traffic without imploding. ATM network protocol can be used since it has shorter data units and built-in QoS for video on demand and VoIP. The QoS priority levels for traffic type are listed as follows:

0 = Best effort
1 = Background
2 = Standard (spare)
3 = Excellent load (business critical)
4 = Controlled load (streaming multimedia)
5 = Voice and video (interactive media and voice) <100 ms latency and jitter
6 = Layer 3 network control reserved traffic <10 ms latency and jitter
7 = Layer 2 network control reserved traffic = lowest latency and jitter

ITU-T G.hn standard provides QoS via contention-free transmission opportunities (CFTXOPs), which allocates the flow and negotiates a contract with the network controller and supports non-QoS operation of contention-based time slots.

Multi Service Access Everywhere defines a QoS concept along with Platforms for Networked Service (PLANETS) to define discrete jitter values including best effort with four QoS classes: two elastic and two inelastic. This concept includes end-to-end delay where the packet loss rate can be predicted. It is easy to implement with simple scheduler and queue length, nodes can be easily verified for compliance, and end users notice the difference in quality. It is primarily based on

the usage of traffic classes, selective call admission control concept, and appropriate network dimensioning.

The Internet is a series of exchange points interconnecting private networks, with its core being owned and managed by a number of different Network Service Providers. The two approaches to QoS in modern packet-switched networks are parameterized and differentiated. The parameterized system is based on the exchange of application requirements within the network using IntServ, RSVP, and a prioritized system where each packet identifies a desired service level to the network. DiffServ implements the prioritized model using marked packets according to the type of service they need. At the IP layer, DSCP markings use the first 6 bits in the TOS field of the IP packet header. At the MAC layer, VLAN IEEE 802.1q and IEEE 802.1D do the same. Cisco IOS NetFlow and the Cisco Class-Based QoS Management Information Base are leveraged in the Cisco network device to obtain visibility into QoS policies and their effectiveness on network traffic. Strong cryptography network protocols such as Secure Sockets Layer, I2P, and VPNs are used to obscure the data being transferred. All electronic commerce on the Internet requires cryptography protocols. This affects the performance of encrypted traffic, which creates an unacceptable hazard for customers. The encrypted traffic is otherwise unable to undergo deep packet inspection for QoS. QoS protocols are not needed in the core network if the margin is high enough to prevent delay. Newer routers are capable of following QoS protocols with no loss of performance. Network providers deliberately erode the quality of best effort traffic to push for higher-priced QoS services. Network studies have shown that adding more bandwidth was the most effective way to provide for QoS.

11.11 Military radios and data links

11.11.1 The joint tactical radio system

The JTRS, sometimes referred to as "Jitters," is a military data link system utilizing SDRs that provide interoperability on the battlefield for voice, data, and video communications. In addition, the system deliver command, control, communications, computers, and intelligence (C4I), include missile guidance command, control, and communications for air-to-air, air-to-ground, and ground-to-ground systems. The JTRS affords the ability to handle multiband, multimode, and multiple channel radios. It was initially designed to cover an operating spectrum of 2 to 2,000 MHz, but this has been expanded to beyond 2,000 MHz to allow higher frequency operation and provide more bandwidths for multiple users and increased data rates.

The main principle behind JTRS is to provide a software programmable radio that is interoperable with all existing legacy radios. In addition, the system needs to provide wideband networking software for mobile ad hoc networks and security or encryption to prevent unwanted users and jammers from entering the network. There are several challenges facing the JTRS.

- Interoperability. There is a need to provide interoperability with different modulation and frequency waveforms for all legacy and future radios on the battlefield. This is a challenge to be compatible with so much diversity in the current communication data links. The RF sections are the most challenging, with components such as filters, antennas, and amplifiers not being able to cover the entire frequency bandwidth. RF integrated circuits are being developed to help mitigate these problems. Tunable filters can be used, but the cost generally increases with complexity. In addition, work is being done to design broadband antennas and antenna switching arrays to cover the wide band of frequencies. The modulation waveform for generation and detection is designed with SDRs that can be software and firmware programed for different modulation/demodulation schemes.
- Antijam. To provide interoperability with multiple radios and data links, the system is required to handle a very wide bandwidth. The wider the bandwidth for operation, the more vulnerable the system is for jammers to affect the signal integrity and performance. Unless these different types of frequencies and modulation schemes are switched from one to another, including their bandwidths, jamming could be a problem.
- Security. Encryption is required to prevent unauthorized users from entering the network. This adds complexity and cost but is necessary to provide the security needed for military operations. One of the major problems is the lack of encryption for each of the legacy radios in the network. The system must be able to handle different encryption schemes or provide a universal encryption system.
- Cost. For several applications, especially for simple controls such as missile guidance, the size and cost of the data link are critical to the customer. With enhanced data link capability, this cost increases, and it may not be practical to use a full capability data link and network protocol. The complexity of the data links used for these types of applications needs to be minimized to avoid size and cost increases.
- Networking. The network must provide interoperability for multiple communication links with different types of radios. Increases in interoperability require networking to avoid collisions of the multiple users. The network protocols have to be designed to handle different legacy radios and future radios and possibly to provide a gateway that can be used for translating different legacy radios. SDRs can be reprogrammed to provide different modulation/demodulation methods, protocols, signal processing, and operational frequencies. They can also be used to provide a network-centric gateway to translate these differences between different radios. In addition, critical networking aspects such as self-forming and self-healing in mesh type networks are utilized to provide highly versatile, high-performance networks. Due to rapid movement on the battlefield coupled with the joint aspect of operations with multiple users from the military, the operational network needs to be dynamic with a constant addition and subtraction of nodes. This network management needs to be real time and versatile to handle the demands and scenarios that are required by the users.

11.11.2 SDRs

SDRs are software programmable radios that can handle different waveforms and multiple modulation/demodulation techniques and can be modified using real-time loaded software in the field or programmed before operation. To extend the use of SDRs, a system with the ability to sense the current waveforms and automatically adjust and adapt to the radios and environment has been developed known as CRs.

SDRs have the capability to program the current waveforms to be interoperable with multiple systems. This is accomplished using DSPs or field-programmable gate arrays (FPGAs) and on-board memory and reprogramming these devices for the waveform and modulation/demodulation processes for the desired radio or data link.

A program that combines legacy and future radios for interoperability is called the Programmable Modular Communications System. It uses SDRs as a key component in allowing software programmable waveform modulation and demodulation, encryption, signal processing, and frequency selection.

11.11.3 Software communications architecture

Software communications architecture (SCA) is essential to the JTRS program and provides a basis for software waveforms. It is a nonproprietary, open systems architecture framework and is required to ensure operability between users. SCA compliance governs the structure and operation of the JTRS using programmable radios to load waveforms, to provide the communication and data link functions, and to allow for networking of multiple users. The SCA standard provides for interoperability among various types of radios by using the same waveform software that can be loaded in multiple users, providing network-centric capabilities. Security is required and becomes a challenge in the overall development of the radios. Type 1 security is generally required for development of the JTRS radios. Software upgradability is also important in the development of the JTRS radios, which allows new waveforms in the future to be compatible with the JTRS system.

11.11.4 JTRS radios (clusters)

The JTRS is designed to be interoperable with legacy data link communication systems. In addition, future growth and capabilities are incorporated to extend the ability of the system to provide new modulation and higher data rate capacities. JTRS development is broken down into five types of radios, each for a particular use and requirements:

- Ground Mobile Radio (GMR) program (Cluster 1): This effort is involved with developing radios for Army and Marine Corps ground vehicles, Air Force Tactical Air Control Parties, and Army rotary-wing applications. This also includes developing a wideband network waveform (WNW), which is the next-generation IP-based waveform to provide ad hoc mobile networking. Recently, Cluster 1 has been revised for ground vehicles only.

- JTRS Enhanced Multi-Band Inter/Intra Team Radio (JEM) (Cluster 2): This effort is involved with upgrading an existing handheld radio by adding JTRS capability to an existing enhanced multiband inter/intrateam handheld radio (MBITR). This is directed by the Special Operations Command (SOCOM). In addition, this radio is SCA compliant. A further development is a JTRS-enhanced MBITR (JEM), which includes rigorous government evaluations and tests.
- Airborne, Maritime, Fixed Site (AMF) program (Clusters 3 and 4): Clusters 3 and 4 have been combined to form this program. This effort incorporates the Army, Air Force, and Navy support. Cluster 3 was the maritime/fixed terminal development for the Navy. Cluster 4 was led by the Air Force and provided Air Force and Naval aviation radios for rotary and fixed-wing aircraft.
- Handheld, Manpack, Small Form Fit (HMS) (Cluster 5): This effort includes handheld radios, manpack radios, and man-portable radios that are small form fit radios for the Army's future combat systems and other platforms such as unattended ground sensors, unattended ground vehicles, unmanned aerial vehicles (UAVs), robotic vehicles, intelligent munitions, and weapon and missile systems. These radios are part of the Soldier Level Integrated Communications Environment (SLICE) program, which utilizes the soldier radio waveform (SRW) and operates in different bands at frequencies of 450 to 1,000 MHz and 350 to 2,700 MHz. Burst data rates are from 450 to 1.2 Mbps and 2 to 23.4 kbps for low probability of intercept.

The Mobile User Objective System (MUOS) is a cellular network that provides support for the HMS radios. This helps in ultra-high frequency Demand Assigned Multiple Access (DAMA) protocols for satellite communications.

A summary of the different JTRS radios is shown in Table 11.5.

11.11.5 JTRS upgrades

Originally planned to span a frequency range of 2 MHz to 2 GHz, the JTRS has now been expanded to frequencies above this top range to satisfy space communications requirements. JTRS HMS radios are used by the individual solder and are designated as the JTRS HMS' AN/PRC-154 Rifleman SDR radios. These include Type 2 Encryption (Phase 2—Type 1) and new SRWs. They are designed to work in ad hoc networks and provide new position tracking capabilities for the individual soldiers. The JTRS HMS' AN/PRC-155 is larger manpack and contains

- SRW
- WNW
- MUOS satellite-communications waveform
- SINCGARS.

The Small Airborne Networking Radio JTRS system is used for helicopter–soldier communication using a SDR for both voice and data.

A list of legacy radios is listed in Table 11.6(a) and (b).

Table 11.5 JTRS radio development

JTRS radios	Users	Characteristics	Features
GMR—ground mobile radio program (Cluster 1)	Army and Marine Corps ground vehicles, Air Force Tactical Air Control Parties, and Army rotary-wing	Next generation IP-based waveform to provide ad-hoc mobile networking	Wideband network waveform WNW Revised for ground vehicles only: ground mobile radio or GMR
JEM—JTRS enhanced multiband inter/intrateam radio (Cluster 2)	Army, Navy, Air Force Special Operations Command (SOCOM)	Upgrade handheld radio by adding JTRS capability to an existing MBITR 30–512 MHz	JTRS-enhanced MBITR or JEM includes rigorous government evaluations and tests
AMF—Airborne, Maritime, Fixed Site program (Clusters 3 & 4)	Air Force and Navy	For rotary and fixed wing remote and maritime vessels	Airborne maritime fixed-station JTRS AMF
HMS—Handheld, Manpack, Small Form Fit (Cluster 5)	Army Soldier Level Integrated Communications Environment (SLICE)	SRW 450–1,000 MHz 350–2,700 MHz Burst data rates 450 kbps–1.2 Mbps LPI 2–23.4 kbps MUOS, DAMA	JTRS Manpack, JTRS HMS, unattended ground sensors and vehicles, UAVs, robotic vehicles, intelligent munitions, weapon and missile systems

Table 11.6(a) Types of legacy radios

Communication link	Frequency of operation	Modulation	Data rate	Features
SINCGARS—single channel ground air radio system	30–87.975 MHz	FM, CPFSK	16 kbps	Single channel (8 possible) or FH (6 hop sets, 2,320 freq/hop set)
EPLRS: enhanced position location reporting system	425.75–446.75 MHz	CP-PSK/TDMA		BW-24 MHz. Burst 825–1,060 ms, 2 ms time slot. FH, 83 MHz channels, 57 kbps VHSIC SIP and 228 kbps VECP
HAVE QUICK II—military aircraft radio UHF used by the Air Force	225–400 MHz	AM/FM/PSK	16 kbps digital voice	FH
UHF SATCOM	225–400 MHz	SOQPSK, BPSK, FSK, CPM	75 bps–56 kbps	MIL-STD-188-181, -182, -183, and -184 protocols, DAMA
WNW—wideband networking waveform	2 MHz to 2 GHz	OFDM, DQPSK	5 Mbps	
Soldier radio and wireless local area network WLAN	1.755–1.850 and 2.450–2,483.5 GHz	TBD	16 kbps voice, 1 Mbps data	IEEE 802.11b, 802.11e, and 802.11g
Link 4A	225–400 MHz		5 kbps	TADIL C
Link 11	2–30 and 225–400 MHz	QPSK	1,364 and 2,250 bps	TADIL A
Link 11B	225–400 MHz	FSK	600, 1,200, and 2,400 bps	TADIL B
Link 16	960–1,215 MHz	MSK	2.4, 16 kbps & data 28.8 kbps to 1.137 Mbps	TADIL J, Ant Diversity, FH, TDMA
Link 22	3–30, 225–400 MHz	MSK	Compatible with Link 16	FH, TDMA
VHF-AM civilian Air Traffic Control	108–137, 118–137 MHz	AM	Analog	BW—25 kHz US, 8.33 kHz European

Table 11.6(b) Types of legacy radios

Communication link	Frequency of operation	Modulation	Data rate	Features
High frequency (HF) ATC data link	1.5–30 MHz	NA	300, 600, 1,200, and 1,800 bps	Independent side band (ISB) with automatic link establish (ALE), and HF (ATC)
VHF/UHF-FM land mobile radio (LMR)	Various, see features	NA	16 kbps	25–54, 72–76, 136–175, 216–225, 380–512, 764–869, 686–960 MHz
SATURN: second generation anti-jam tactical UHF radio	225–400 MHz	PSK	NA	
IFF—identification friend or foe. IFF/ADS/TCAS will support data at	1,030 & 1,090 MHz	NA	689.7 bps	
DWTS—digital wideband transmission system ship-board system	1,350–1,850 MHz	NA	Multiple from 144 to 2,304 kbps	
Cellular radio and PCS	824–894, 890–960, & 850–1,990 MHz	Various, see features	10 kbps, 144/384 kbps, and 2 Mbps	TR-45.1 AMPS, IS-54 TDMA, IS-95b CDMA, IS-136HS TDMA and GSM and 3GSM, 2.5 G, 3G, W-CDMA, and CDMA2000
Mobile satellite service (MSS)	1.61–2 [2.5] GHz	NA	2.4 to 9.6 kbps	Includes VHF and UHF MSS bands, LEOs & MEOs, Iridium, Globalstar, and others

Integrated broadcast service module (IBS-M)	225–400 MHz	NA	2.4, 4.8, 9.6, and 19.2 kbps	
BOWMAN—UK Tri-Service HF,VHF and UHF tactical communications system	NA	NA	16-kbps digital voice	
AN/PRC-117G©—man-portable radio platform	30 MHz–2 GHz	NA	Up to 5 Mbps	Embedded, programable, INFOSEC capabilities
VHF FM	30–88 MHz	FM	16 kbps	
VHF AM ATC (extended)	108–156 MHz	AM	16 kbps	
VHF ATC data link (NEXCOM)	118–137 MHz	D8PSK	4.8 kbps Voice 31.5-kbps data	
UHF AM/FM PSK	225–400 MHz & 225–450 MHz	AM/FM, PSK	16 kbps	
MUOS	240–320 MHz	NA	2.4, 9.6, 16, 32, & 64 kbps	Support the common air interface

11.11.6 JTRS network challenge

The initial concept for the JTRS was to have all of the waveforms and networking protocols available at each JTRS radio. To do this requires high-performance DSPs, FPGAs, and a lot of memory. Generally, the main obstacles are antennas and RF circuits to handle the vast differences in frequencies and bandwidth. This requires size, cost, complexity, and power. This may be acceptable for some applications, but small size, weight, and power requirements for portable radios, man-pack radios, and weapon and missile systems are critical.

11.11.7 Gateway and network configurations

A possible alternative to this concept is to use a gateway, which provides all of this capability and can be used to translate different signals for different radio users. The gateway has the network control, using a network system that connects all of the different users to a single point, known as a star topology, and also employs TCP and IP (Figure 11.4).

Another type of network, called a bus topology network, allows users to communicate directly with each other without requiring a gateway to relay the information (Figure 11.5). All of the users are connected to a bus, and according to various protocols to avoid sending data at the same time, they are able to communicate with everyone else who is connected. In addition, the bus topology can include multiple gateways, linked together in the bus network, and the users of the bus can send and receive their data to and from the gateway. The gateway then translates and sends the data to other users connected to the bus.

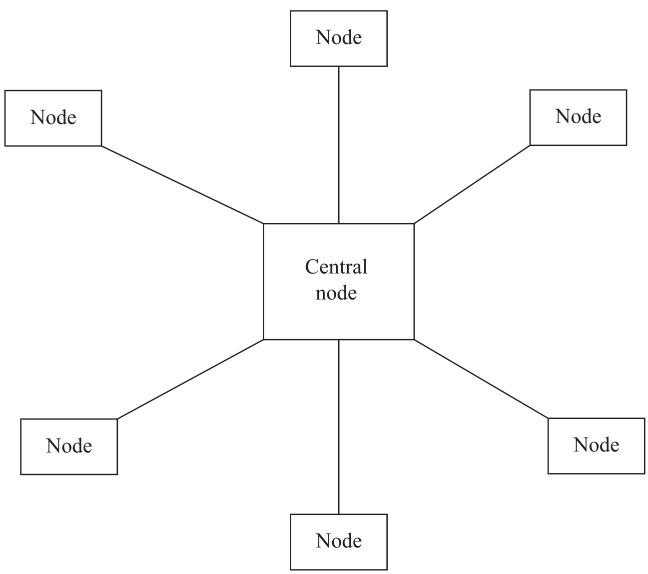


Figure 11.4 Star topology network

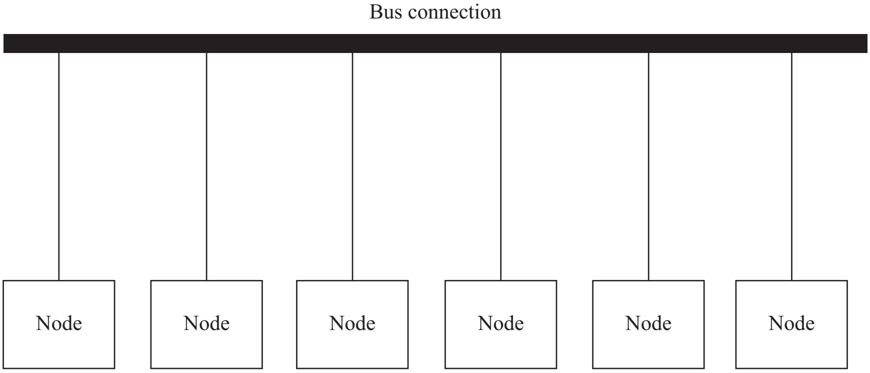


Figure 11.5 Bus topology network

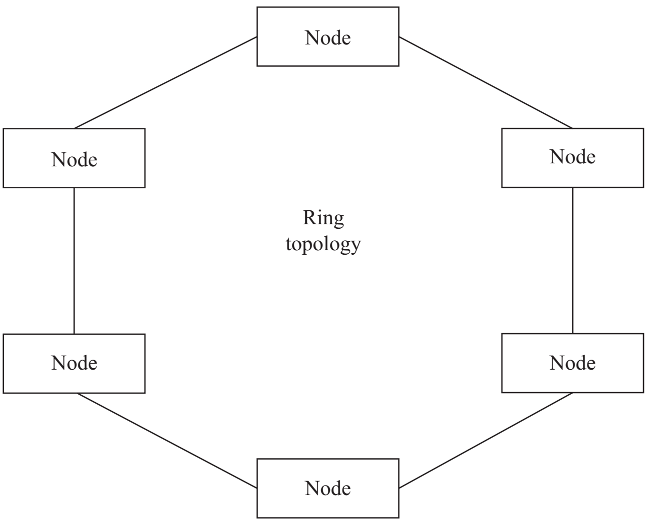


Figure 11.6 Ring topology network

A ring network connects all of the users in a ring (Figure 11.6), and the data are passed around it, going from user to user. If the protocol is set up to receive the sent message, then the data are received. Here again, the ring network can include multiple gateways, and the users on the ring can send and receive their data to and from the gateway, and the gateway would translate and send the data to other users connected to the ring.

A high-performance network used in many applications is known as a mesh network. It is used in wireless applications for both commercial and military networks. The main advantage of mesh networks is their ability to establish and maintain network continuity and route signals and links in the presence of

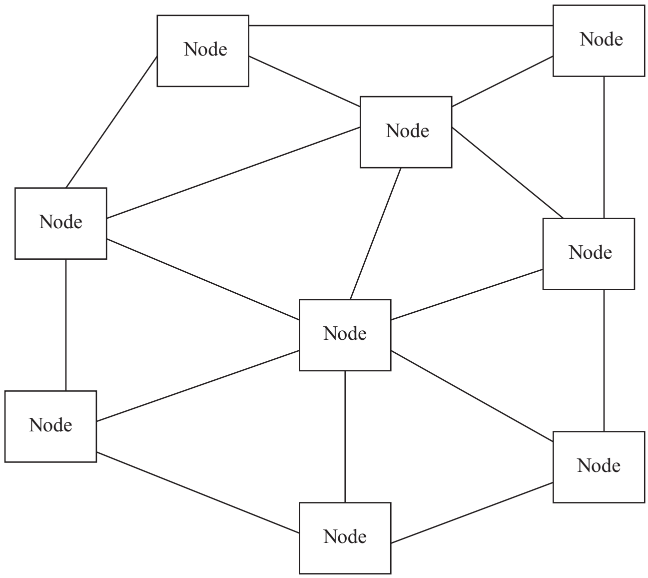


Figure 11.7 Wireless mesh network

interference and node dropouts in the network. They are able to maintain network integrity and perform real-time repair of the network during a failure without shutting down. This is known as self-forming and self-healing. They do not contain a central hub that controls each of the users; instead, the data links are established directly between users, which reduce single point failures (Figure 11.7).

Wireless mesh networks require minimum system administration, maintenance, and support. They are very useful in establishing networks for broadband communications. They overcome inherent connectivity limitations and provide automatic configuration and routing.

Whatever type of network is used, the main purpose is to provide an overall system to communicate with many different radios and links into one big communication network. One of the steps to take in the near future is to develop interoperability—for example, making the SINCGARs radio interoperable with the HAVEQUICK II radio. From there, the system must expand and grow by adding other legacy radios, and as some of the legacy radios become obsolete, they can be replaced with SDRs.

LAN standards. Standards have established to help set the direction and interoperability of different manufacturer's hardware, and they are continually being updated and new standards established to keep up with the new technologies, speed, and applications. The Institute of Electrical and Electronic Engineers (IEEE) established a standards committee known as the LAN Standards Committee 802. This group created the 802.3 standard for bus topologies using CSMA/CD and the 802.5 standard dealing with token rings.

Ethernet. Ethernet uses a CSMA/CD scheme, in which a node senses if the bus is clear, then sends out a transmission, and monitors if there is another node trying to transmit. If it detects another transmission, then it sends out a signal telling all nodes to cease transmission. The other nodes cease for a random period of time and then try again if the channel is clear. This helps to prevent an endless loop of collisions between nodes trying to use the bus. Propagation delay can also increase the probability of collision. Someone may be using the system, and the other node may not have received his transmission due to the propagation delay so the node attempts to use the system.

There are four common standards using Ethernet: 10BASE-5, 10BASE-2, 10BASE-T, 100BASE-T, and 1000BASE-T. The designation tells the user the physical characteristics. The 10 describes the operating speeds (10 Mbps), the BASE stands for baseband, the 5 is distance (500 m without a repeater), and the T stands for unshielded twisted pair (UTP). 100BASE-T provides 100 Mbps data rates under the IEEE 802.3u standard using CSMA/CD technology. The 100BASE-TX is a version that provides 100 Mbps over two pair of CAT-5 UTP or two pair of Type 1 shielded twisted pair (STP). If other types of cables are used, then the designation changes. For example, fiber cable uses X, 1000BASE-X; coax cable uses CX, 1000BASE-CX. Other terms for 1,000 Mbps or 1 Gbps Ethernet are GbE or 1GigE.

Network layers. In the discussion of networking, several layers of the design have been established to help in designing interoperable systems and to use different layers with other designs. The standard for network layers is the Open System Interconnection Seven-Layer Model. Some of these layers may not be included in the design of a network system, but the following gives an overview of the different types and basically what they do. More detailed description can be found in other references:

1. Physical layer. The lowest level network layer and is mainly concerned with voltages, currents, impedances, power, signal waveforms, and also connections or wiring such as the RS-232 serial interface standard.
2. Data link layer. Used to communicate between the primary and secondary nodes of a network. Involved with activating, maintaining, and deactivating the data link. Also concerned with framing the data and error correction and detection methods.
3. Network layer. Specifies the network configuration and defines the mechanism where messages are divided into data packets.
4. Transport layer. Controls end-to-end integrity, interface or dividing line between the technological aspects and the applications aspects.
5. Session layer. Network availability, log-on, log-off, access, buffer storage, determines the type of dialog, simplex, duplex.
6. Presentation layer. Coding, encryption, compression.
7. Application layer. Communicates with the users program. Controls the sequence of activities, sequence of events, general manager.

A chart showing the different layers is given in Figure 11.8.

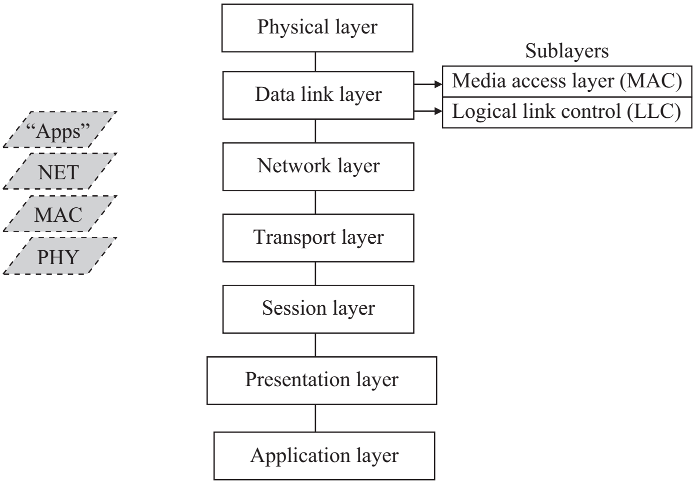


Figure 11.8 OSI seven network-layer model

11.11.8 Link 16

Link 16 (TADIL J) uses J messages and has been looked at to provide interoperability and networking in a complete battlefield system. This discussion addresses Link 16 Class 2, Tactical Data Link for C3I.

The Joint Tactical Information Distribution System (JTIDS) uses Link 16 as its core. The Multifunctional Information Distribution System (MIDS), with its reduced size, is used in the F/A-18 aircraft. The MIDS provides interoperability between the MIDS and JTIDS. The Army is using JTIDS with a rate of less than 8 kbps and employs a combinational interface of the Position Location Reporting System and JTIDS hybrid interface.

In addition, the MIDS radio has been upgraded to be compatible and interoperable with JTRS requirements and is referred to as MIDS JTRS.

Link 16 is a multiuser system with low duty cycle, pseudo-randomly distributed in the frequency code domain on a slot-by-slot basis. The range is 300 nm nominal, with the ability to extend the range to 500 nm. Any Link 16 terminal can operate as a relay to further increase the range of the system. The power output is 200 W low power and 1,000 W high power if a high-power amplifier is added to the system. The power can be reduced by eliminating both the voice channels and tactical air navigation (TACAN).

The system operates at least two antennas, for antenna diversity, to prevent multipath and jamming. The antennas used are broadband vertical antennas in the ultra-high frequency (L-band) 960 to 1,215-MHz operating frequency range. Special filtering is added to prevent interference with the TACAN, distance measuring equipment, and identification friend or foe (IFF) by using a dual-notch band-pass filter, which eliminates IFF at 1,030 and 1,090 MHz. The frequency of operation of the Link 16 system is in two bands: ultra-high frequency, 300–1,000 MHz, and L-band, allocated 960–1,215 MHz (969–1,206 MHz) with an IF of 75 MHz.

11.11.9 Link 16 modulation

The modulation of Link 16 is minimum shift keying (MSK). A single message consists of 5 data bits, which is spread at a 5-MHz rate using 32-bit cyclic code shift keying (CCSK) to produce an approximately 8-dB processing gain: 10 log(32/5). Each of the 5 bits shifts the 32-bit sequence 3 as follows:

1st bit = 0, no shift of the 32 bit code: 000110 . . . .
2nd bit = 1, 1 shifts the 32 bit code to the left: 001100 0
3rd bit = 2, 2 shifts to the left 011000 00

The 32-bit signal is XOR with the 32 chips of the PN-sequence. It uses the final 32 chip signal to modulate the frequency shift using differential—no change fl, change fh. The demodulation lines up the PN code and strips of the code. It uses correlators for the 32-bit sequence to determine the 5 data word.

The CCSK Link 16 also uses FH, with each message pulse on a different frequency per pseudo-noise (PN) code, with up to 20 different hop PN codes to prevent mutual interference. Therefore, each hop frequency contains one message, which contains 5 data bits spread by 32 MSK chips. The message pulse is 6.4 μs on and 6.6 μs off, for a total of 13 μs.

The data rates are variable, with rates of 28.8, 31.6, 57.6, and 115.2 kbps, with a capacity up to 238 kbps, which can be expanded to 2 Mbps. In addition to data, the Link 16 provides two digital voice channels with rates of 2.4 and 16 kbps.

In addition, Link 16 provides FEC. This is a Reed-Solomon (RS) (31,15). The data word is 15 5-bit characters (70 bits plus 5 bits parity) and is expanded to 31 RS characters (155 bits). If there is a need to receive more than half bits error free, then the entire message can be regenerated. FEC and interleaving across hops have the ability to correct channel bits that are lost, but the voice channel contains no FEC.

An important part of Link 16 is encryption. The system uses COMSEC/TRANSEC, encrypted by KGV-8B. In addition, it is encrypted using FH and jitter (the random time period the terminal waits to enter system).

11.11.10 TDMA

Hopped MSK pulses are assigned a time slot for TDMA, up to 60 hops/time slot at a rate of 76.923 kHps, or 1/13 μs. The data link hops both the data and the sync pulse. The frame is 12 s, with 1,536 time slots/frame. A total of 24 h is divided into 112.5 epochs, which is equal to 12.8 min each. The epochs are divided into 98,304 time slots, which equals 7.8125 ms/time slot. Access to Link 16 rotates among all users every 12.8 min (98,304 slots). Each cycle is an epoch, and assignments are made against a 12-s repeating frame.

11.11.11 "Stacked" nets

Stacked nets are networks that allow multiple users to occupy the same time slots because the users have different PN codes (20 total) selected for their hop sequence and some process gain due to MSK.

11.11.12 Time slot reallocation

Time slot reallocation (TSR) was developed for the Navy and is similar to a DAMA system but without a central control station. To avoid collisions and contention access, the user waits for the next time slot if one time slot is busy.

11.11.13 Bit/message structure

The first pulse is the start of the time slot, called jitter, followed by the synch pulse, then the header timing pulses containing the following information:

ID
Address classification
FEC instructions
RS FEC for the header is (16,7) and interleaved

The message can be single or double pulse. The standard is double pulse, with the same data in each pulse. There are additional variations to this standard. The time allocated for each part of the message is as follows:

Time slot jitter pulse = 2.418 ms
Time propagation allowed in the design = 2.040 ms
Time for the data plus header = 3.354 ms
In extended slot, jitter time is used for data; total data time = 5.772 ms

A summary of the Link 16 performance specifications is shown in Table 11.7.

Link 16 has been the preferred method of communication in the military arena. The challenge is for industry to generate a network design centered on this technology and determine if it is feasible to use this data link for all applications, or perhaps a subset, or possibly a totally new method for accomplishing interoperability for C4I.

Table 11.7 Link 16 performance specifications

RF system	**Frequency**	**Modulation**	**Data rate**	**Other networking**
Link 16	300–1,000 MHz 960–1,215 MHz (969–1,206 MHz)	Spread spectrum MSK 32 chips/5 bits FH 1 hop/message LPI, antijam TDMA FEC Encryption Double pulse	31.6, 57.6, and 115.2 kbps 238 kbps capacity, expandable to 2 Mbps 200 W low power 1,000 W high power with HPA	Stacked nets Time slot reallocation TSR 20 different PN codes

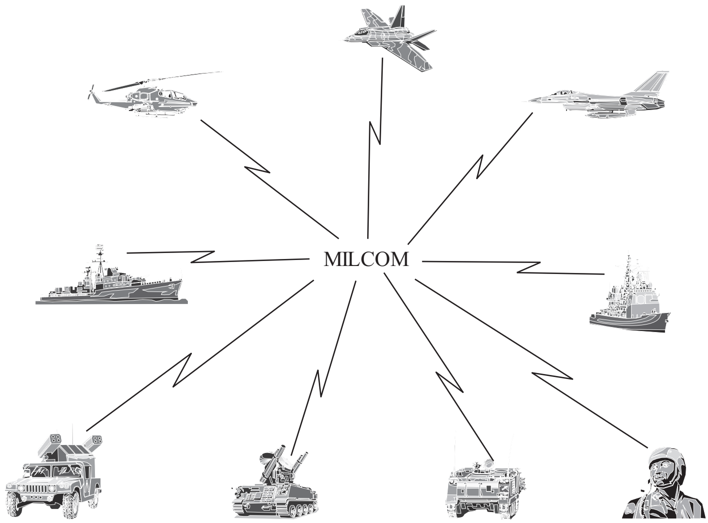


Figure 11.9 Military communications network for communications, command, control, and weapon systems

11.12 Summary

Broadband and home networking will shape the future. New standards are being reviewed as new technologies are developed and as the data rates increase. With several incoming signals to a home, such as voice, data, and video, there is a need to provide optimal distribution throughout to allow for easy access.

Commercial and military communications are ubiquitous and constantly growing and improving using new technologies. Networking is becoming important on the military battlefield, and JTRS and Link 16 play important roles in the interoperability of communication devices. The development of these and other new technologies will provide the military with a network for all communication devices present and in the future (Figure 11.9).

11.13 Problems

1. What are the three mediums used in broadband distribution in the home?
2. What is a common technique used in power-line communications to maximize the frequency band with multiple users?
3. What criteria are used to determine if two signals are orthogonal?
4. Describe how the desired signal can be retrieved among other signals in an OFDM system.
5. What is a major drawback of home PNA?
6. What are the three main challenges that face the JTRS?

7. How do SDRs help to solve the interoperability problem?
8. What are the four standard network topologies?

Further reading

Core Specification of the Bluetooth System Specification Volume 1, Version 1.1 February 22 2001.

Lough, Daniel L., T. Keith Blankenship, and Kevin J. Krizman. 2007. “A Short Tutorial on Wireless LANs and IEEE 802.11.” http://decweb.bournemouth.ac.uk/staff/pchatzimisios/research/IEEE%20802_11%20Short%20Tutorial.htm.

Profiles Specification of the Bluetooth System Specification Volume 2, Version 1.1 February 22 2001.

Chapter 12
Satellite communications

Satellite communications are becoming a viable means of providing a wide range of applications for both the commercial and military sectors. The infrastructure for distributing signals covers the widest range of communications methods; even the most remote places on earth like the South Pole can have communications via satellite. Satellite's bandwidth, field of view, and availability, along with combining this technology with other types of communications systems, produce a ubiquitous infrastructure that provides communications worldwide.

12.1 Communications satellites

Communications satellites are sent into space in a geostationary orbit and are made up of a space platform and the payload. The payload is the equipment and devices mounted on the space platform. One satellite may contain multiple payloads for multiple applications, including different data links operating with their own frequencies and antennas, which are all mounted to the space platform and sent into space.

Each satellite is equipped with attitude control to ensure that the directional antennas, which are attached on the spacecraft, are all pointed toward the earth and that the antenna beam is focused on the intended area on the earth (Figure 12.1).

Several types of disturbances change both the attitude of the geostationary satellite and the orbit of the satellite by a slight amount. The satellite's attitude stabilizers compensate for the changes in attitude in order to keep it pointing toward the earth. The orbit of the satellite is changed due to movements in the north and south plane in a figure-eight pattern over a 24-h period.

12.2 General satellite operation

End users are connected to an earth station at their location, whether it is a personal system or a network-serving multiple users. This earth station provides two-way communications between it and the satellite, referred to as the space station. The space station relays the information to another earth station that provides the services, whether it is data links; military operations for command, communications, and control; Internet services; telephone; fax; video; music; or other types of information (Figure 12.2).

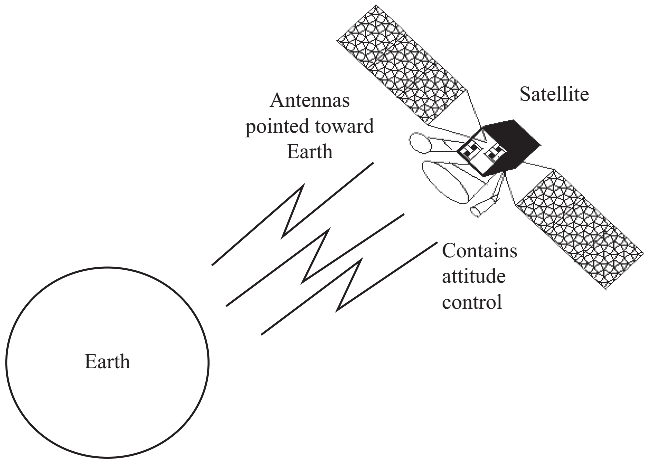


Figure 12.1 Attitude control to ensure antennas pointing direction

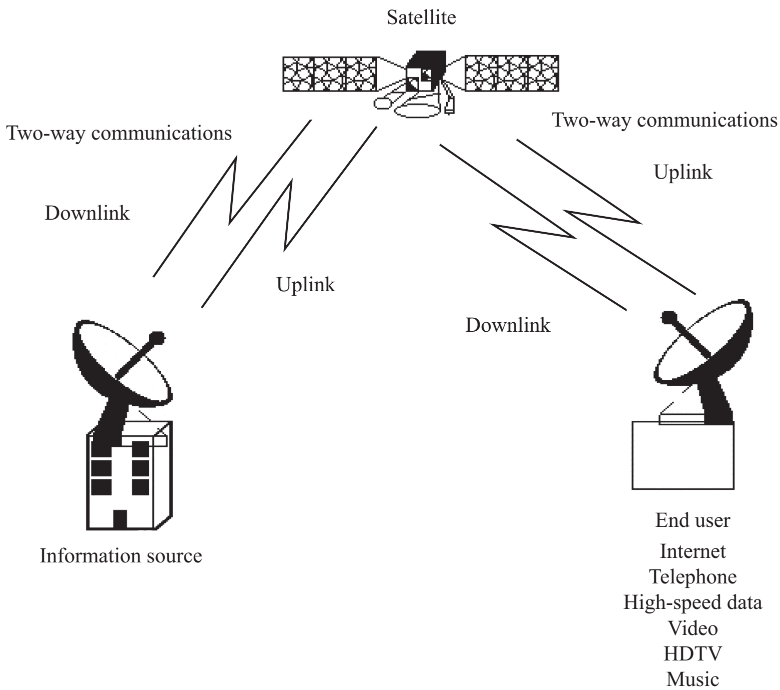


Figure 12.2 Satellite operational communications link

12.2.1 Operational frequencies

Five different bands of operation are used for satellite communications. The lowest band is called the L-band, which operates with an uplink at 1.6 GHz and a downlink at 1.5 GHz using a narrow bandwidth. The next band is the C-band, which operates at around 6 GHz for the uplink to the satellite and 4 GHz for the downlink from the satellite to the ground station. The next band, which is generally used by the military, is the X-band, which operates at around 8 GHz for the uplink and 7 GHz for the downlink. The next band, which has become popular for telecommunications, is the Ku-band, which operates at around 14 GHz for the uplink and 11–12 GHz for the downlink. The most widely used of these four are the C-band and the Ku-band.

A fifth band, called the Ka-band, is becoming popular for broadband communications and military applications. It operates at 30 GHz for the uplink and 20 GHz for the downlink and provides a much higher bandwidth for high-speed data and more simultaneous end users. A summary of the different frequency bands including very small aperture terminal (VSAT) frequency bands is shown in Table 12.1.

These are approximate frequencies and represent bands currently in use. In addition to the five standard satellite bands mentioned previously, very high frequency and ultrahigh frequency VHF/UHF satellite bands are also being used for lower data rates and operate across multiple frequencies from 136 to 400 MHz. Another satellite communications band that is emerging is the Q-band which operates in the 33 to 50-GHz extremely high frequency range. Bands are continually expanding and changing, so these numbers may not be exact and may vary slightly. Also, along with frequency changes, bandwidths change for different systems and uses. In future, other bands will eventually open up for satellite communications as the demand increases.

Table 12.1 Standard and VSAT satellite frequency bands of operation

VSAT satellite frequency bands		
Frequency band	**Uplink (GHz) Earth station to satellite**	**Downlink (GHz) Satellite to earth station**
C-band	5.925–6.425	3.700–4.200
Extended C-band	6.725–7.025	4.500–4.800
Ku-band	14.000–14.500	10.950–11.700

Standard satellite frequency bands

Band	Freq (GHz)	Uplink (GHz)	Downlink (GHz)	Other
L	1.6/1.5	1.6	1.5	Narrow BW
C	6/4	5.850–6.425	3.625–4.200	500 MHz BW
X	8/7	7.925–8.425	7.250–7.750	500 MHz BW mostly military
Ku	14/12	14–14.5	11.5–12.75	500 MHz BW
Ka	30/20	27.5–31	17.7–21.2	3.5 GHz BW

12.2.2 Modulation

The main modulation scheme for the Intelsat/Eutelsat time division multiple access (TDMA) systems is coherent quadrature phase-shift keying (QPSK). Coherent QPSK uses four phase states providing 2 bits of information for each phase state. Since this is coherent QPSK, the absolute phase states 0, 90, 180, and 270 are used to send the information. The data rate is approximately equal to 120 Mbps using a bandwidth of 80 MHz. The Intelsat system uses 6/4 GHz (6 GHz uplink, 4 GHz downlink), while the Eutelsat system uses 14/11 GHz. The radio frequency is downconverted to typical intermediate frequencies including 70 MHz, 140 MHz, and 1 GHz.

Techniques have been developed to increase the efficiency of the satellite link. One of these techniques is called digital speech interpolation (DSI), in which data signals are transmitted during the dead times of voice channels or telephone calls. This provides a method of sending data and utilizing times when voice is not being sent. Another technique to increase link efficiency combines DSI with a decrease in voice speed from 64 to 32 kbps; this is called digital circuit multiplication equipment.

12.2.3 Adaptive differential pulse-code modulation

One of the best and most efficient ways to quantize analog signals is via adaptive differential pulse-code modulation (ADPCM). The analog signals are sampled at a rate greater than the Nyquist rate. Each of the samples represents a code value for that sample, similar to an analog-to-digital converter. The differential part of the ADPCM describes a process by which the sampled value is compared with the previous sampled value and measures the difference. The adaptive part of the modulation scheme means that the step size can be made finer or coarser depending on the difference measured. This helps in tracking large analog voltage excursions. If the samples are continually increasing, then the step value is increased. If the samples are continually decreasing, then the step size is decreased (Figure 12.3).

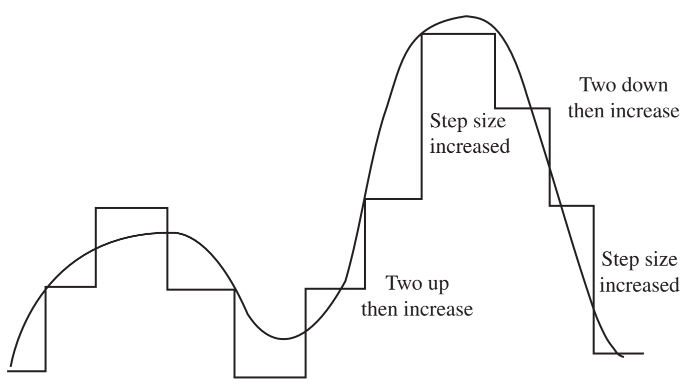


Figure 12.3 ADPCM to convert analog-to-digital signals

12.3 Fixed satellite service

Satellite communications using geostationary satellites to provide services to the end user are known as fixed-satellite service (FSS). FSS is a radio communications service between fixed points on the earth using one or more geostationary satellites.

12.4 Geosynchronous and geostationary orbits

Most satellites are approximately 22,000 miles above the earth's surface. If they follow a geosynchronous orbit, then the period of the orbit is equal to the earth's rotation. If they are only geosynchronous, then they are not in a fixed position as far as the view from the earth. The orbits are inclined with respect to an orbit around the equator (Figure 12.4).

Most communications satellites follow a geostationary orbit. These orbits need to be geosynchronous but should travel in a circular orbit on the equatorial plane. This allows satellites to be in a fixed position in relation to the view from the earth. Therefore, they follow a circular orbit on the equatorial plane, circling the earth once every 24 h, the same time it takes the earth to rotate once on its axis. By synchronizing to the earth's rotation, so that the satellite follows the earth at approximately the same speed as the angular rotation of the earth, the satellite will look stationary to a fixed point on the earth (Figure 12.4).

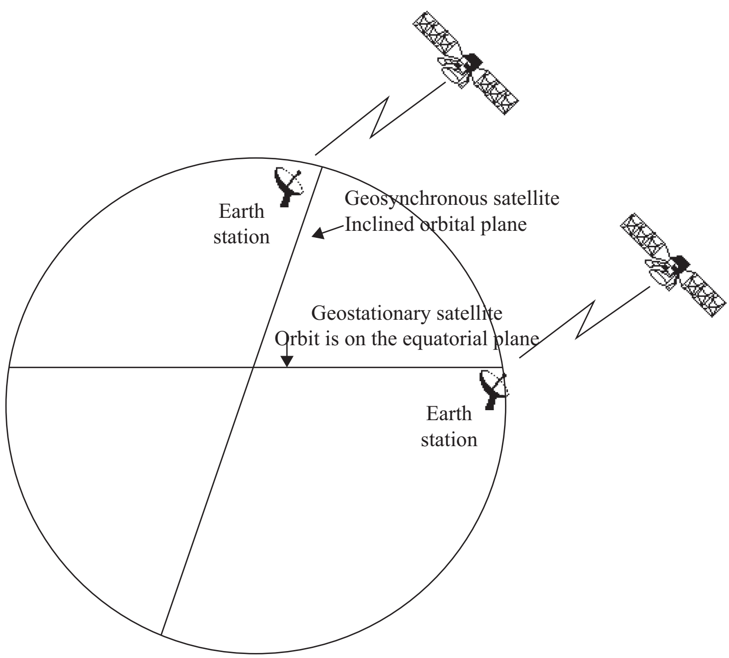


Figure 12.4 Geosynchronous and geostationary satellite orbit

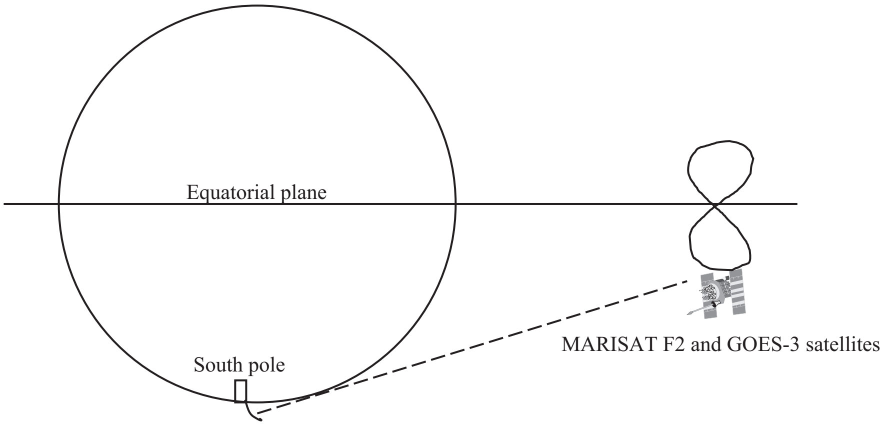


Figure 12.5 Satellite tilt orbit in a figure-8 pattern

This provides a continuous link with a given satellite at any time during the day or night. Therefore, the communications link will be available all the time unless there is a problem with the data link between the earth and the satellite. Another advantage is that the ground equipment becomes less expensive and easier to operate, since tracking a reasonably stationary satellite is much easier than tracking a satellite with high angular velocity.

Even though geostationary satellites appear stationary, they actually drift in an orbital figure-eight pattern north and south of the equator (Figure 12.5). Satellites are designed to help mitigate this drift so that the earth stations can be stationary without automatic tracking controls. After they have been in orbit for close to their space lives, this figure-eight pattern becomes much larger. For 30-year-old satellites, the orbital tilt becomes large. Because of this, satellites such as the Marisat F2 and GOES-3 were used for communications to remote places such as the South Pole. From the South Pole, it is impossible to see satellites on the equatorial plane. But because these older satellites follow this figure-eight pattern when they travel south of their orbit, they become visible from the South Pole which allows communications to other parts of the world (Figure 12.5).

12.5 Ground station antennas

Even though the satellite appears to be stationary for a geostationary orbit, large antennas with very narrow beam widths generally require an automatic tracking device to provide the best performance. Because smaller antennas have small apertures and wide beams, they generally do not require a tracking device. Digital satellite systems for television use a small antenna and do not require any tracking device. The antenna is installed at the site and usually does not have to be adjusted for several years.

The gain of a ground station antenna is largely dependent on the diameter of the antenna and is also frequency related. The equation for the gain of the parabolic antenna that is often used for satellite communications systems is:

$$G_t = 10 \log\left[n(\pi(D)/\lambda)^2\right]$$

where n is the efficiency factor <1, D is the diameter of the parabolic dish, and λ is the wavelength.

Low-cost systems operating in the Ku- and Ka-bands using small antennas approximately 1–2 m in diameter are known as VSATs. VSATs provide two-way communications to a central location called a hub. They are used mainly for businesses, schools, and remote areas. They basically connect remote computers and data equipment to the hub via satellites.

Three types of antenna systems are incorporated in satellite communications:

- Primary focus antenna system. The feed is positioned in front of the primary reflector, and the signal is reflected once from the feed to the intended direction of radiation. The single reflector antenna system is generally less expensive and provides the simplest design (Figure 12.6(a)).
- Cassegrain antenna system. This antenna system uses a dual-reflector arrangement. The feed comes from the back of the primary reflector and sends the signal to a convex subreflector, which is mounted in front of the primary reflector. Then, the signal is fed to the primary reflector where it is reflected in the desired direction (Figure 12.6(b)). Cassegrain antennas are more efficient than primary focus antennas for several reasons, the main one that the subreflector can be adjusted or formed to optimize the signal reflection and to focus less energy toward the blockage of the subreflector, which falls in the path of the primary reflector. They are also easier to maintain, since the feed horn is located at the base of the reflector, which provides easy access compared with having the feed horn on structures in front of the reflector.
- Gregorian antenna system. This antenna system is also a dual-reflector system, very similar to the Cassegrain antenna. The main difference is that the subreflector is concave instead of convex. The feed is mounted on the rear of the primary reflector and sends the signal to the subreflector. The signal is reflected in the concave subreflector, which causes a crossover of the reflected signal. The reflected signal is sent to the primary reflector, where it is reflected in the desired direction (Figure 12.6(c)). Because of this crossover, the subreflector must be a greater distance from the primary reflector.

12.6 Noise and the low-noise amplifier

The low-noise amplifier (LNA) is critical to the design of a satellite communications system. The system performance is improved directly for each dB of improvement in the LNA. The LNA is the main element that sets the noise figure for the satellite receiver unless there are large losses after the LNA or the

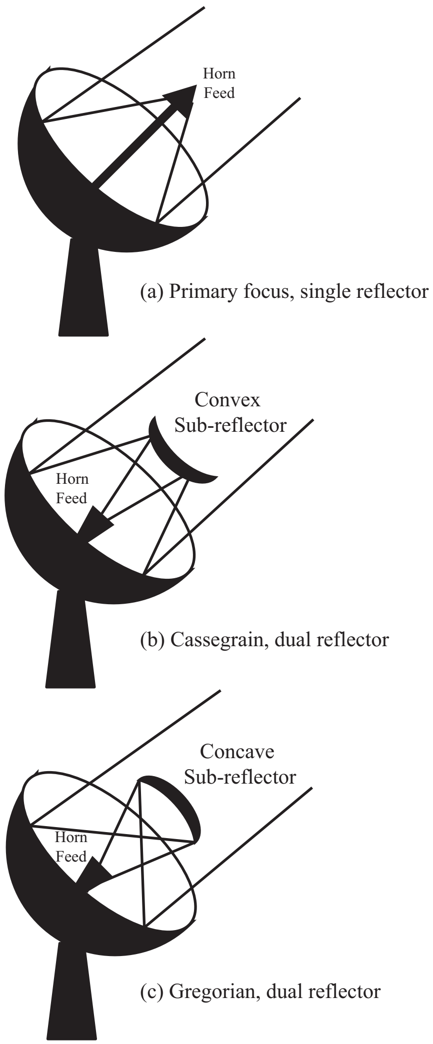


Figure 12.6 Types of antennas for satellite communication systems. (a) Primary focus, single reflector, (b) Cassegrain, dual reflector, and (c) Gregorian, dual reflector

bandwidth becomes larger, which may degrade the receiver performance significantly. A complete noise analysis is addressed in Chapter 1. Another method of doing link budgets that is commonly used for satellite communications is evaluating the system using equivalent noise temperature. Most link budgets for terrestrial systems use noise power, signal power, and the standard *kTBF* at the output of the LNA, with the losses from the antenna affecting the signal level in the link on a one-for-one basis. These are the standard receiver equations and

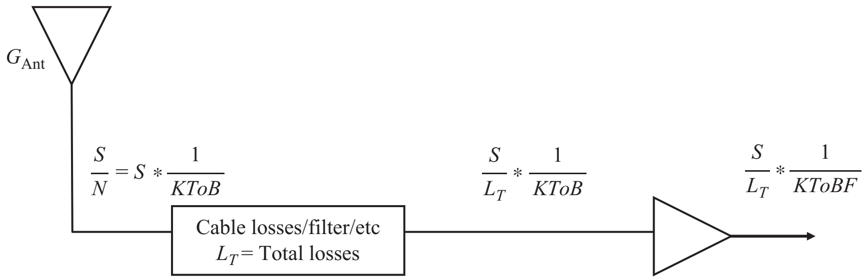


Figure 12.7 Standard receiver equations for signal and noise levels

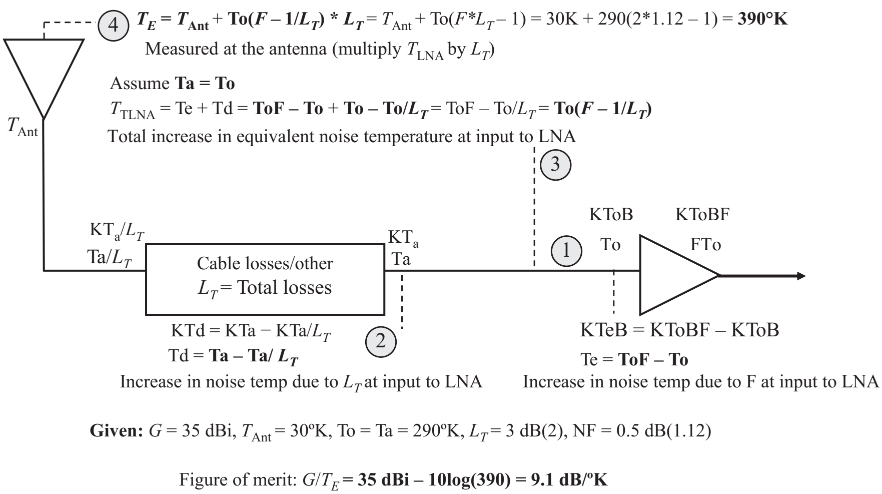


Figure 12.8 Equivalent noise temperature analysis, G/T

methods for determining the signal and noise levels in the receiver (Figure 12.7). However, for satellite transmission systems, the analysis uses equivalent temperatures and converts the noise factor of the LNA to an equivalent temperature (Figure 12.8):

$$T_e = (F - 1)T_o$$

where T_e is the equivalent temperature due to the increase in noise of the LNA noise factor, F is the noise factor of the LNA [noise factor = 10log(noise factor)], and T_o is the 290 K.

Working in reverse to determine the noise factor:

$$T_e = (F - 1)T_o = T_oF - T_o$$
$$F = (T_e + T_o)/T_o$$

Therefore, the noise factor, which is the increase in noise for the receiver, is equal to the additional temperature (T_e) added in front of an ideal receiver that produces the same amount of increased noise. So using this equation, if the noise figure of the LNA is low, say 0.5 dB, then the noise factor is 1.12, and the equivalent temperature T_e is equal to 34.8 K. To illustrate this, the noise power at the output of the LNA is equal to kT_oF if the bandwidth and gain is ignored along with Boltzmann's constant. The input noise power of the LNA is equal to kT_o with the same assumptions. Thus, the increase in noise power due to the noise factor of the LNA is equal to:

$$= N_i = kT_oF - kT_o$$

Since the equivalent noise increase is desired for the link budget, the equivalent noise T_e is solved by eliminating the k:

$$T_e = T_oF - T_o = (F - 1)T_o = (1.12 - 1)290 = 34.8\,\mathrm{K}$$

Since the satellite link budget is in terms of temperature, the losses in the system between the antenna and the LNA need to be converted into temperature (Figure 12.8). This is done by taking the difference in noise power due to the total losses:

$$\text{Noise power difference of attenuator} = N_d = kT_a - kT_a/L_T$$
$$\text{Temperature difference} = T_d = T_a - T_a/L_T = T_a(1 - 1/L_T)$$

where T_d is the temperature difference due to losses or attenuation, L_T is the total losses, and T_a is the temperature of the losses or attenuation, generally equal to $T_o = 290$ K.

Note that if $L_T = 1$, there are no losses and $T_d = 0$. As L_T approaches infinity, then $T_d = T_a = T_o$. Therefore, the total equivalent temperature out of the antenna for the receiver (T_r) is:

$$T_r = T_e + T_d = (F - 1)T_o + T_o - T_o/L_T = T_oF - T_o/L_T = T_o(F - 1/L_T)$$

assuming T_o is the temperature of the losses.

Using the previous example and 3 dB (2) of cable losses and solving for T_r:

$$T_r = T_o(F - 1/L_T) = 290(1.12 - 1/2) = 180\,\mathrm{K}$$

This is the equivalent temperature measured at the input to the LNA.

To find the total equivalent temperature for a system measured at the input to the LNA, the noise due to the antenna (T_{Ant}) needs to be included. The antenna noise is attenuated by the cable loss (L_T), and the total equivalent temperate at the LNA would be:

$$T_{\mathrm{Total\ LNA}} = T_{\mathrm{Ant}}/L_T + T_o(F - 1/L_T)$$

The equivalent temperature at the antenna input is calculated by integrating the gain and noise temperature in the direction of the antenna. This is accomplished

by multiplying the previous equation by the cable loss between the LNA and the antenna to provide the temperature equivalent at the location of the antenna. The noise of the antenna is calculated as follows:

$$T_E = L_T(T_{\text{Ant}}/L_T) + T_o(F - 1/L_T) = T_{\text{Ant}} + T_o(L_T \times F - 1)$$

This is the total equivalent noise temperature at the antenna (Figure 12.8).

Once the temperature of the antenna is calculated, then a figure of merit, which is used to evaluate different satellite systems, can be used to determine the quality of the receiver. The figure of merit is equal to the gain of the antenna divided by the total temperature at the antenna (*G*/*T*), which is usually given as dB/K. Since the gain of the antenna is usually expressed in dBi, *G*/*T* is calculated by

$$G/T = G\ \text{dBi} - 10\log(T_E)$$

A typical value is around 30 dB/K, but this value will vary tremendously with the size of the antenna, frequency, LNA, losses, and other factors.

12.7 The link budget

The link budget uses the noise and LNA along with the signal level to determine the range and quality of the satellite link. To calculate the signal or carrier level at the input to the receiver, an effective isotropic radiated power (EIRP) is determined from the transmitter.

12.7.1 EIRP

The EIRP out of the transmitting antenna is the power output of the power amplifier (PA), losses from the PA through the antenna, and the gain of the antenna. The power out of the PA is usually expressed in dBm (10log(mW)) or dBW (10log(W)).

Several types of PAs are used for transmitting signals to the satellite, including traveling wave tubes (TWTs), Klystrons, low-cost field-effect transistor (FET) amplifiers, and gallium nitride (GaN) amplifiers. TWTs offer wide bandwidths, 500 MHz and higher, good group delay, and the ability to handle many signal inputs. Klystrons are expensive, are high-power only, and have narrow bandwidths in the 40 to 80 MHz range. The FET solid-state (SS) amplifiers offer low-cost solutions and include many of the features of the other technologies, though generally not for extremely high-power outputs. However, GaN SS amplifiers provide high-power outputs and are being used more, since their cost and complexity is lower. A general power requirement for different types of signals includes 1 W per channel for telephone signals and 1 KW per television carrier.

The losses are in dB and are subtracted from the PA output power; the antenna gain is in dB and is added to the final result. The EIRP for the satellite is calculated as follows:

$$\text{EIRP} = P_t + L_{ta} + G_{ta}$$

where P_t is the power output of the PA (in dBm), L_{ta} is the total attenuation in the transmitter (in dB), and G_{ta} is the gain of the transmitter antenna (in dB).

12.7.2 Propagation losses

The propagating channel is calculated for the satellite data link according to the losses discussed in Chapter 1, with the major contributor being free-space loss:

$$A_{fs} = 20\log[\lambda/(4\pi R)] = 20\log[c/(4\pi Rf)]$$

where λ is the wavelength, R is the slant range, same units as λ, c is the speed of light, 300×10^6 m/s, R is in meters, and f is the frequency.

The free-space loss for a satellite system operating in the X-band is approximately 200 dB. Depending on the conditions in the atmosphere, other losses such as clouds, rain, and humidity need to be included in the link budget. These values are usually obtained from curves and vary from day to day and from region to region. Each application is dependent on the location, and a nominal loss, generally not worst case, is used for the link analysis.

12.7.3 Received power at the receiver

Other losses include multipath loss, which for satellite systems is generally small if the antennas are sited properly. For low-angle satellites, multipath can become a factor. In addition, the receiver also contains losses, similar to the transmitter, which need to be included in the link budget. Finally, implementation and spreading losses are included. The received power at the LNA is:

$$P_s = \mathrm{EIRP} + A_{fs} + L_p + L_{ra} + G_r + L_{\mathrm{multi}} + L_{\mathrm{imp}} + L_{ss}$$

where EIRP is the effective radiated power with respect to an isotropic radiator (in dBW), A_{fs} is the free-space loss, L_p is the propagation loss, L_{ra} is the receiver losses, G_r is the gain of the receiver antenna, L_{multi} is the multipath losses, L_{imp} is the implementation losses due to hardware, and L_{ss} is the spreading losses due to spread spectrum.

A typical link budget for satellite communications is provided to evaluate the trade-offs of the system and to ensure that the receivers on either end have enough signal-to-noise ratio for reliable communications (Table 12.2).

12.7.4 Carrier power/equivalent temperature

Often in satellite systems, the power received by the receiver is called the carrier power (C), which is also compared with an equivalent temperature (C/T). A simplified form of combining the losses, dividing both sides by temperature, and converting all values to actual values (not dB) is:

$$C/T = \mathrm{EIRP} \times G_r/(L_T \times T) = G_r/T \times \mathrm{EIRP}/L_T$$

where C/T is the carrier power/equivalent temperature, G_r/T is the figure of merit, EIRP is the effective isotropic radiated power, and L_T is the total losses from the EIRP of the transmitter to the receiver.

Power flux density (PFD) is used to determine the amount of power radiated by the antenna in a direction at a large distance per unit of surface area. For an

Table 12.2 Link budget for the uplink from the earth station to the satellite

Link budget analysis:				
	Slant range (km)	**Freq. (GHz)**	**Power (W)**	
Enter constants.........	35,000	6	10	
Enter inputs...............	**Inputs**	**Power levels**		**Temp**
Transmitter	Gain/loss (dB)	Sig. (dBm)	Noise (dBm)	Kelvin
Tran. Pwr (dBm) =		40		
Trans. line loss =	–0.5	40		
Other (switches)	–0.5	39		
Trans Ant gain =	47	86		
Ant. losses =*	–1	85		
EIRP		85		
Channel				
Free-space loss =	–198.89	–113.89		
Rain loss =	–0.20	–114.09		
Cloud loss =	–0.10	–114.19		
Atmloss (etc) =	–0.10	–114.29		
Multipath loss =	–2.00	–116.29		
Receiver				
RxAnt gain =	47.00	–69.29		
G/T dB/K				**25.20**
Total noise temp at Ant.				152
Antenna noise TA				30
Ant. losses =	–0.02	–69.31		
Other (switches)	–0.25	–69.56		
Rec. line loss =	–0.25	–69.81		
Total losses LT	–0.52			
RF BW (MHz)**	100.00		–94.00	
Total noise temp. at rec.				134
Equiv. Temp. Te =				75
LNA noise Fig. =	1.00		–93.00	
LNA gain =	25.00	–44.81	–68	
LNA levels =		–44.81	–68.00	
Receiver gain =	60.00	15.19	–8.00	
Imp. loss	–4.00	11.19		
Detector BW	10.00		–18	
Det. levels =		11.19	–18	
S/N =		**29.19**		
Req. Eb/No				
Req. Eb/No		12	Pe=10exp-8	
Coding gain =	4.00	8.00		
Eb/No margin =		**21.19**		

isotropic radiator, the equation equals

$$\mathrm{PFD} = P/\left[4(\pi)d^2\right]$$

where PFD is the power flux density, P is the output power, and d is the distance.

Add the gain to this to find the PFD for a gain antenna.

With satellite's increased power and more sensitive receivers, earth stations are becoming less costly and smaller. For example, the Intelsat V satellite contains 50 transponders on-board and operates with just 5–10 W of power. The total bandwidth for the Intelsat V is 500 MHz to provide high-speed data for multiple users.

12.8 Multiple channels in the same frequency band

To utilize the band more efficiently and to provide more data or users, two different schemes are employed. The first scheme uses beam pointing at two different points on Earth using the same satellite. Each beam is focused on an area of the earth so that the same antenna can be used for two systems in the same band and frequency. The narrower beamwidth illuminates less coverage on the earth's surface. However, this provides more EIRP so that the earth stations can be less expensive.

Another way to increase the data capacity for a given band is polarization. If orthogonal polarizations are used for two different channels, they can use the same antenna and bandwidth with minimal interference between them. In theory, horizontal and vertical polarizations are orthogonal, as are left-hand circular polarization and right-hand circular polarization. Cross polarization can occur, which degrades the separation of two channels that are orthogonally polarized. Often orthogonal polarization is used to provide increased isolation between the transmitter and the receiver for transceiver operation.

For a system to use polarization for frequency reuse in a satellite communications system, the isolation between the channels should be at least 25 to 30 dB. The main causes that degrade the isolation are the Faraday effect (Earth's magnetic field) and atmospheric effects (rain or ice crystals). Also, multipath can alter the polarization during reflection of the signals on a surface, and propagation through the troposphere or ionosphere can lead to polarization disturbances.

12.9 Multiple access schemes

Multiple access is important in satellite communications to allow multiple users in the same bandwidth, satellite, and antenna system. Two basic methods are used to provide for multiple users. The first is a multiplexing scheme called preassigned multiple access (PAMA), which permanently assigns a user to a channel or time. An example of PAMA is using time slots and assigning a given time slot to each end user. Since they are permanently assigned, the users will have that given time

slot and will be multiplexed with all of the other users, similar to time division multiplexing used in other communications links.

The other method of multiplexing is called demand assigned multiple access (DAMA). This method is a true multiple access scheme similar to TDMA, which is used in other communications links. This system is used on an "as needed" basis, and each user takes any available time slot when needed.

Another method of providing for multiple users is frequency division multiplexing or frequency division multiple access (FDMA). These methods use frequency division to handle multiple users on one band. In systems that incorporate FDMA, the channels are separated using different frequency slots for different users. Other techniques to provide for multiple users are beam focusing, so that each of the beams covers a different area of the earth from the satellite, and antenna polarization as mentioned earlier.

12.10 Propagation delay

One of the problems with real-time communications using satellites is the large distance between the earth station and the satellite. There is an approximate propagation delay of 275 ms. The two-way propagation delay is double that time, or 550 ms. Since this propagation delay is so long, echo cancelation is vital to the quality of the communications link, especially for voice applications. Video applications, especially one-way broadcasting, generally does not need echo cancelation techniques.

12.11 Cost for use of the satellites

The cost is determined by the type of transmission, type of signal sent, and the length of the transmission time. The types of transmissions include PAMA, DAMA, and occasional. PAMA is the most costly, since it basically ties up an entire multiple access slot all of the time. Demand systems such as DAMA are less costly, since the access slot is assigned only when it is in use. The least costly is the occasional use, which is billed only for the time that it is used.

The types of signals sent using the satellite link include voice, video, and data. Many systems do not include voice as an option, with the Internet being the most important. For more information on the cost of use of satellites, see the Intelsat tariff handbook, which specifies detailed charges.

12.12 Regulations

The International Telecommunication Union ITU radio regulations provide the rules for satellite communications to avoid interference and confusion. The main specifications supplied by the ITU include frequency band allocations, power output limitations from either the earth station or the satellite, minimum angles of elevation for earth station operation, and pointing accuracy of antennas.

12.13 Types of satellites used for communications

The Inmarsat satellite system's main use is for maritime communications for ships and shore earth stations. Currently, it is being employed for broadband communications and extended coverage. The satellites cover the main bodies of water, but earth stations do not have to be located on the shore as long as the satellites are visible. The Inmarsat satellites operate on C- and L-bands. The FSS gateway for telephone service is via the Inmarsat satellites.

Another satellite system is the Intelsat system. Intelsat provides broadband communications, including digital data, video, telephones, and other communications. There are several types of Intelsat satellites in use today.

Other satellites that cover various sections of the world are Eutelsat, which are used mainly for European countries, and PanAmSat, which are used for Central and South American countries. Each satellite's coverage needs to be evaluated for optimal performance, satellites in view, usage, and cost.

With the push toward providing broadband communications, including high-speed data and Internet connections, many companies are developing both Ku- and Ka-band satellite communication systems. They use geosynchronous earth orbit satellites, low earth orbit satellites (LEOS), and hybrid satellite constellations to provide both Ka- and Ku-band operation, with data capacities ranging from 30 Gbps to systems providing much greater data capacities for the end users.

LEOS are broken down into two subclasses: little and big. The orbits of little LEOS are from 90 min to 2 h and are less than 2,000 km high. They can be used for approximately 20 min per orbit. They operate in the following frequency bands listed:

137–138 MHz
400.15–401 MHz
148–149.9 MHz

Some little LEOS follow a polar orbit at an altitude of 850 km and are sun synchronous, which means that the satellite passes over a certain area at the same time every day.

Big LEOS include satellite systems such as Iridium, Teledesic, and others.

Another class of satellites is medium earth orbit satellites. They orbit at an altitude of approximately 10,000 km, with a 12-h orbit, which means that these satellites travel around the earth twice a day. They are generally 55° inclined from the equatorial plane. These types of satellites include GPS.

Geostationary satellites were discussed at length earlier in this chapter. They orbit at an altitude of approximately 35,800 km.

And finally, there are highly elliptical orbit satellites. Since their orbits are elliptical, the perigee is approximately 500 km, and the apogee is approximately 50,000 km. Their orbit time is between 8 and 24 h for one revolution around the earth.

Handheld phones have been developed including the Iridium phone (upgraded from the first phones) and IsatPhone Pro phone that works with the Inmarsat satellite. Both are small handheld devices that are available commercially.

12.14 System design for satellite communications

The main design criteria for satellites are the transponder bandwidth, EIRP, and *G*/*T*. Typical figure of merit values for 4 GHz range from 41 dB/K using a 30-m antenna using a parametric amplifier to 23 dB/K for a 4.5-m antenna using an FET amplifier. For a space station receiver at 6 GHz, typical figure of merit values ranges from 19 dB/K using an FET transistor LNA providing a wide coverage area using a wide beamwidth to −3 dB/K for a pencil beam antenna. For the earth station, the main design criteria are *G*/*T*, antenna gain, system noise temperature, and transmitted power. The overall system design criteria are operational frequency bands, modulation methods, multiple access parameters, system costs, channel capacity, and overall system performance.

12.15 Summary

Satellites are now providing extensive coverage for communications and data link operations in remote areas. The satellite connection consists of a remote earth station, a satellite, and another earth station. This triangle forms a two-way communication link to provide the remote earth station with access to all types of communications, including data links, military operations, Internet, video, voice, and data at high data rates.

Satellite and ground systems use mainly five bands: L, C, X, Ku, and Ka. The latter is becoming popular for both commercial and military sectors. A geostationary orbit is used so that the ground station tracks a fairly stationary transceiver and the satellite appears to be stationary.

A link budget is used to determine the power, gains, and losses in a system and also determines the figure of merit. Multiple access schemes are used to allow multiple users on the same band. Costs are associated with the type of system used and the length of use.

12.16 Problems

1. What system has the most ubiquitous infrastructure for communication systems?
2. What is the difference between a geostationary satellite and a geosynchronous satellite?
3. What phenomenon of a satellite orbit allows communication with the South Pole?

Further reading

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Chapter 13
Global navigation satellite system

The last few years have shown an increased interest in the commercialization of the global navigation satellite system (GNSS), which is often referred to as the global positioning system (GPS) in various applications. A GPS system uses spread spectrum signals—binary phase-shift keying (BPSK)—emitted from satellites in space for position and time determinations. Until recently, the use of GPS was essentially reserved for military use. Now there is great interest in using GPS for navigation of commercial aircraft. The US Federal Aviation Administration implemented a wide area augmentation system for air navigation to cover the whole United States with one system. There are also applications in the automotive industry, surveying, and personal and recreational uses. Due to the increase in popularity of GPS, and since it is a spread spectrum communication system using BPSK, a brief introduction is included in this text.

13.1 Satellite transmissions

The NAVSTAR GPS satellite transmits a direct sequence BPSK signal at a rate of 1.023 Mbps using a code length of 1,023 bits. The time between code repetitions is 1 ms. This is known as the coarse acquisition (C/A) code, which is used by the military for acquisition of a much longer precision code (P-code). Commercial industries use the C/A code for the majority of their applications. There are 36 different C/A codes that are used with GPS, and they are generated by modulo-2 adding the C/A code with a different delayed version of the same C/A code (Figure 13.1). Therefore, there are 36 different time delays to generate the 36 different C/A codes. Approximately 30 codes are used presently for satellite operation; each is assigned a pseudorandom noise code number for identification. The space vehicle identification number is also used for identification.

Two frequencies are transmitted by the satellite. The C/A code uses only one of these frequencies, L1, for transmission. The frequency of L1 is 1.57542 GHz. Therefore, there are 1,540 carrier cycles of L1 for each C/A chip. The frequency of L2 is 1.2276 GHz, and it does not contain the C/A code. The P-code, which is used by the military, is transmitted on both frequencies. The P-code is transmitted in quadrature (90°) with the C/A code on L1. The carrier phase noise for 10 Hz one-sided noise bandwidth (B) is 0.1 radian root mean square for both carriers, and the in-band spurious signals are less than -40 dBc.

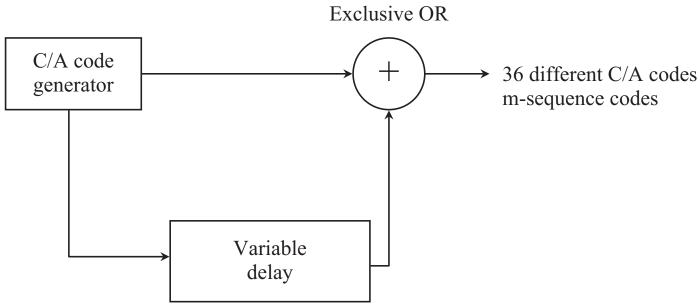

Figure 13.1 Generating 36 different C/A codes

A summary chart is provided for both the C/A code and the P-code as follows:

- C/A code:
  - BPSK spread spectrum modulation waveform
  - 1.023-Mbps chipping rate
  - Code length = 1,023 bits
  - Code repetition rate = 1 ms
  - C/A code 3 dB higher than P-code
  - 36 different C/A codes
  - Sliding correlator demodulation
  - Frequency L1 = 1.57542 GHz, 1,540 cycles/chip
- Military uses precision code (P-code):
  - BPSK spread spectrum modulation waveform
  - 10.23-Mbps chipping rate
  - Code length approximately 1 week long
  - Uses both L1 = 1.57542 GHz and L2 = 1.2276 GHz
  - P-code transmitted in quadrature with C/A code on L1

The antenna used in the satellite for transmissions has a 3-dBi gain and is linear, right-hand circularly polarized. The group delay deviation for the transmitter is within ±3 ns, 2σ.

13.2 Data signal structure

The satellite navigation data are sent out at a 50-bps rate. As a general rule, the information bandwidth should be at least 1 kHz. The data are sent out in 25 full frames with 5 subframes in each frame. Each subframe is 300 bits long (10 words at 30 bits each) providing frames that are 1,500 (5×300) bits long. The total number of bits required to send out the total satellite information is 37.5 kbits ($25 \times 1{,}500$). Therefore, the time it takes to send out the total number of bits at a 50-bps rate is 12.5 min:

$$37.5 \text{ kbits}/50 \text{ bps} = 750 \text{ s} = 12.5 \text{ min}$$

The following five subframes contain the data necessary for GPS operation:

- Subframe 1: Space vehicle (SV) clock corrections.
- Subframes 2 and 3: Complete SV ephemeris data. Ephemeris data contain such things as velocity, acceleration, and detailed orbit definitions for each satellite.
- Subframes 4 and 5: Subframes 4 and 5 accumulated for all 25 frames provide almanac data for 1 to 32 satellites. The almanac data are all the position data for all the satellites in orbit.

Note that the clock corrections and SV or satellite ephemeris data are updated every 30 seconds.

Each of the subframes has an identification (ID) code 001,010,011,100,101, and also uses a parity check, which involves simply adding the number of "1"s sent. If an odd number is sent, the parity is a "1," and if an even number is sent, then the parity is a "0."

13.3 GPS receiver

The standard GPS receiver is designed to receive the C/A coded signal with a signal input of -130 dBm. Since the C/A code amplitude is 3 dB higher than the P-code amplitude and the frequency of the C/A code is 1/10 of the P-code, then using a filter the receiver is able to detect the C/A code messages. Note that the C/A code also lags the P-code by 90° to keep the codes orthogonal and further separated. For most systems, the SV needs to be greater than 5° elevation to keep multipath and jamming signals at a minimum. Since the GPS signal is a continuous BPSK signal, a sliding correlator can be used to strip off the 1.023-MHz chipping signal, leaving the 50-Hz data rate signal.

13.3.1 GPS process gain

Although the received signal is very small, approximately -130 dBm, the data rate, bandwidth, and the noise are small:

Data rate = 50 bps
BW < 50 Hz
Noise < -155 dBm

The chip rate is equal to 1.023 Mcps for the C/A code and 10.23 Mcps for the P-code so that the process gain for each code is equal to:

$$\mathrm{Gp} = 10\log\,(1.023\ \mathrm{Mcps}/50\ \mathrm{bps}) = 43.1\ \mathrm{dB\ for\ C/A\ code}$$
$$\mathrm{Gp} = 10\log\,(10.23\ \mathrm{Mcps}/50\ \mathrm{bps}) = 53.1\ \mathrm{dB\ for\ P\text{-}code}$$

13.3.2 Positioning calculations

The GPS position is calculated using three satellite signals, each of which is received by the GPS unit, which calculates the time difference corresponding to range, called a pseudorange. For each time measured from the satellite to the GPS

receiver, corresponding pseudorange solutions generate a sphere around the satellite (Figure 13.2). Using two satellites and measuring the time or pseudorange solutions for each satellite, these two spheres intersect producing pseudorange solutions around a circle (Figure 13.2). Using three satellites, the three intersecting spheres produce two points that are the only two possible solutions. Since one of the solutions will not be located on the earth's surface, the other solution determines the position of the user (Figure 13.2). The equations for the position of the user with three satellites is therefore

$$\begin{aligned}
\mathbf{PR1} &= \sqrt{(\boldsymbol{x1} - \boldsymbol{xp})^2 + (\boldsymbol{y1} - \boldsymbol{yp})^2 + (\boldsymbol{z1} - \boldsymbol{zp})^2} \\
\mathbf{PR2} &= \sqrt{(\boldsymbol{x2} - \boldsymbol{xp})^2 + (\boldsymbol{y2} - \boldsymbol{yp})^2 + (\boldsymbol{z2} - \boldsymbol{zp})^2} \\
\mathbf{PR3} &= \sqrt{(\boldsymbol{x3} - \boldsymbol{xp})^2 + (\boldsymbol{y3} - \boldsymbol{yp})^2 + (\boldsymbol{z3} - \boldsymbol{zp})^2}
\end{aligned}$$

With three equations and three unknowns, these equations solve the xp, yp, zp position of the user.

To eliminate a clock bias that exists between the satellites and the GPS user, another variable needs to be solved, and it requires one more satellite and one more

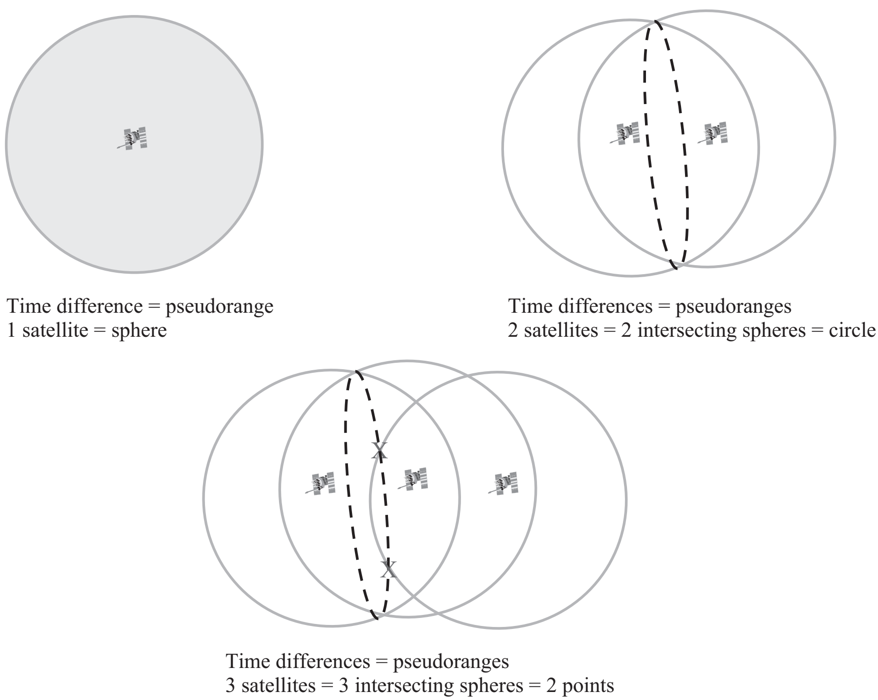

Figure 13.2 Intersecting spheres from each satellite produces the position solution

equation (four equations and four unknowns):

$$\mathbf{PR1} = \sqrt{(\boldsymbol{x}\mathbf{1} - \boldsymbol{xp})^2 + (\boldsymbol{y}\mathbf{1} - \boldsymbol{yp})^2 + (\boldsymbol{z}\mathbf{1} - \boldsymbol{zp})^2} + \boldsymbol{C}_{\mathbf{bias}}$$
$$\mathbf{PR2} = \sqrt{(\boldsymbol{x}\mathbf{2} - \boldsymbol{xp})^2 + (\boldsymbol{y}\mathbf{2} - \boldsymbol{yp})^2 + (\boldsymbol{z}\mathbf{2} - \boldsymbol{zp})^2} + \boldsymbol{C}_{\mathbf{bias}}$$
$$\mathbf{PR3} = \sqrt{(\boldsymbol{x}\mathbf{3} - \boldsymbol{xp})^2 + (\boldsymbol{y}\mathbf{3} - \boldsymbol{yp})^2 + (\boldsymbol{z}\mathbf{3} - \boldsymbol{zp})^2} + \boldsymbol{C}_{\mathbf{bias}}$$
$$\mathbf{PR4} = \sqrt{(\boldsymbol{x}\mathbf{4} - \boldsymbol{xp})^2 + (\boldsymbol{y}\mathbf{4} - \boldsymbol{yp})^2 + (\boldsymbol{z}\mathbf{4} - \boldsymbol{zp})^2} + \boldsymbol{C}_{\mathbf{bias}}$$

Since all of the GPS satellites are synchronized with GPS master time and are also corrected as needed, the clock bias is assumed to be equal.

13.4 Atmospheric errors

The atmospheric path loss for the satellite link is approximately 2 dB. The troposphere and ionosphere cause variable delays, distorting the time of arrival and position (errors are dependent on atmosphere, angle, and time of day). These delays are extreme for low-elevation satellites. To compensate for the errors, ionosphere corrections are implemented using both frequencies (f_{L1}, f_{L2}):

$$\mathrm{PR} = \frac{f_{L2}}{f_{L2}^2 - f_{L1}^2}(\mathrm{PR}(f_{L2}) - \mathrm{PR}(f_{L1}))$$

where PR is the compensated pseudorange, f_{L2} is the frequency of $L2$, f_{L1} is the frequency of $L1$, PR(f_{L2}) is the pseudorange of $L2$, and PR(f_{L1}) is the pseudorange of $L1$.

If the pseudoranges of $L1$ and $L2$ are not known, then the measured and predicted graphs over time need to be used. This would be the case for a C/A code-only receiver. A simple C/A code receiver uses the ionospheric correction data sent by the satellites for a coarse ionosphere correction solution.

13.5 Multipath errors

The angular accuracy for platform pointing is affected by multipath:

$$\sigma_\theta = \sigma_R / L$$

where σ_θ is the angular accuracy of platform pointing, σ_R is the range difference caused by multipath, and L is the baseline.

Good antenna design can reduce multipath to approximately 1 m. The receiver antenna is designed to reduce ground multipath by ensuring that the gain of the antenna is low toward the ground and other potential multipath sources compared to the gain toward the satellites of interest.

Using integrated Doppler (or carrier phase) can also improve the effects of multipath on the overall positioning solution to help smooth the code solution.

Further discussion on smoothing the code solution with the integrated Doppler of the carrier follows later in this chapter.

Another way to reduce multipath errors is to first determine that the error is caused by multipath and then enter it as a state variable in a Kalman filter. This solution requires that the multipath can be measured accurately and is constant.

It is interesting to note that multipath errors can actually give you a closer range measurement. By definition, multipath arrives later than the direct path; however, the resultant autocorrelation peak can be distorted such that the tracking loop (e.g., early–late gate) produces a solution that is earlier than the direct path.

13.6 Narrow correlator

Multipath can cause errors in range due to distortion of the autocorrelation peak in the tracking process. One way to reduce errors is to narrow the correlation peak, which is known as the narrow correlator. The early–late and tau-dither loops are generally operated on half-chips (i.e., a half-chip early and a half-chip late), which provides points on the crosscorrelation peak halfway down from the top on both sides. By using less than half-chip dithering, the points are closer to the peak, reducing the ambiguity caused by multipath (Figure 13.3). The accuracy of the measurement is also greatly improved for the same reasons.

When using the narrow correlator, it is important to note that the precorrelation bandwidth is larger depending on the step size of the correlator. For example,

2 MHz is required for 1 C/A chip ($\pm$[0.5 chip]) standard correlator.
8 MHz is required for 0.1 C/A chip ($\pm$[0.05 chip]) narrow correlator.

The narrower the correlator, the wider the precorrelation bandwidth needs to be because the peak needs to be sharper; thus the input code needs to have sharper rise times to create this sharper peak. If the peak is rounded, then the narrow correlator does not improve performance because of the ambiguity of the samples. One of the drawbacks of the narrow correlator is a larger precorrelation bandwidth (8 MHz compared with 2 MHz), which is also more vulnerable to jammers and noise, and a faster clock is required. However, in many precision navigation applications, this technology is preferred.

For P-code receivers, the P-code null-to-null bandwidth is 20 MHz. Therefore, for the narrow correlator to work, the bandwidth should be approximately 80 MHz, approximately four times, to ensure that the peak of the correlator is sharp. The problem is that the satellite filters this output signal using a bandwidth of approximately 20 MHz. The increased bandwidth in the receiver does not produce sharper peaks in the autocorrelation peak because the high-frequency components are filtered out at the satellite transmitter and therefore the correlation peak is rounded (Figure 13.4). This makes it hard to detect differences in the values that drive the locked conditions.

Another locking mechanism that aligns the correlation peak for the narrow correlator is called a delay-locked loop (DLL). DLL processing can be

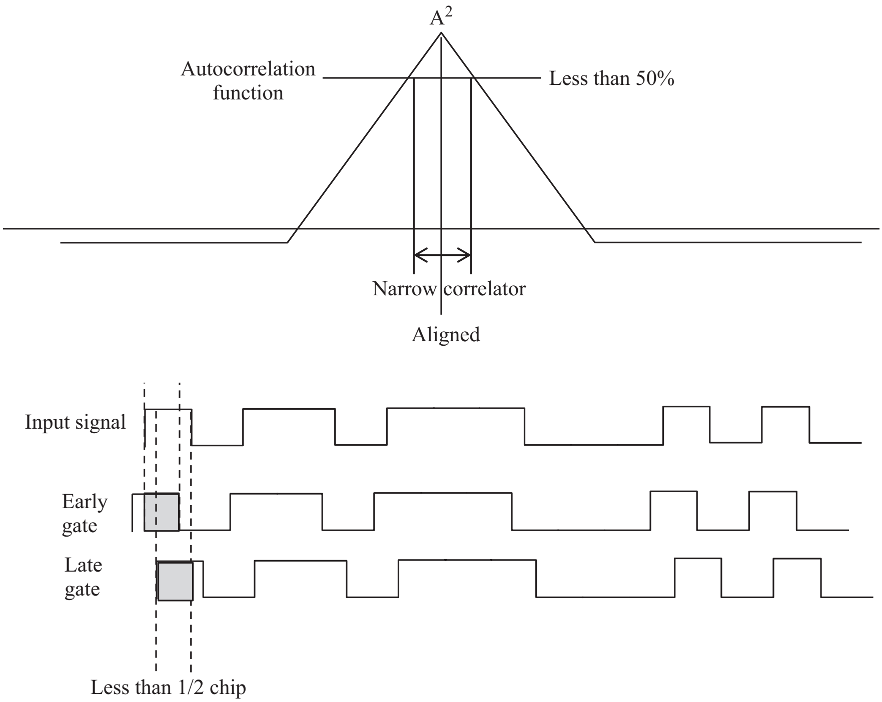

Figure 13.3 Narrow correlator for improved performance over standard correlator

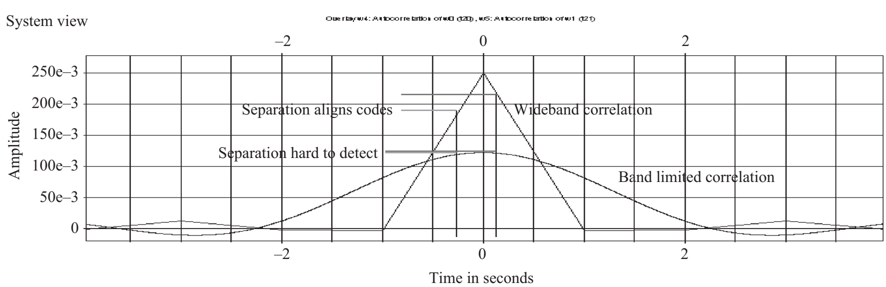

Figure 13.4 Limitations of the narrow correlator

accomplished by using the difference between the punctual correlation value and the early-minus-late value which generates the error off of the punctual. For low signal-to-noise ratios, this may prove advantageous since the punctual signal is higher in amplitude than the early or late gate.

13.7 Selective availability

Selective availability (SA) was developed by the US Department of Defense to make C/A code positioning less accurate. SA is caused by jittering or dithering the code clock and provides an improper description of the satellite orbit, that is, ephemeris data corruption. This clock dithering reduces the accuracy of the C/A code receiver by changing the receive time of the codes. With SA turned on, the worst-case accuracy of a C/A code receiver is approximately 100 m, which also includes other error sources. SA was turned off in May 1, 2000, and is currently not a factor in the accuracy of the GPS C/A code receiver. This improves positioning accuracy from approximately 100 m worst case to approximately 20 m worst case for stand-alone GPS systems. The actual accuracy of the GPS C/A code receiver is dependent on several other parameters and conditions.

Note that many of the specifications written for accuracy are done with SA turned off. This provides a good baseline for comparison, but caution needs to be used when using the actual accuracy numbers if SA is ever turned back on again. The probability of SA being turned back on is remote since there are techniques to cancel SA for accurate C/A code receivers; these will be discussed later in this chapter.

13.8 Carrier smoothed code

The code solution by itself is noisy, particularly with the effects of multipath, SA, and atmosphere. This makes it difficult to obtain accurate measurements for the code-derived pseudoranges and introduces variations in the measurements. To aid in the code solution, the carrier is used to help smooth the variations in the code solution.

The carrier solution is a low-noise solution. It contains a wavelength ambiguity so that it does not know which of the wavelengths it is supposed to be comparing for the phase offset. For example, the difference in phase may be 10°, but it may be off several cycles or wavelengths—for example, 370°, 730°. By using the code solution that contains no wavelength ambiguity and the carrier phase solution, a carrier smoothed code can be produced. The phase change at selected repeatable points of the carrier is used to smooth the noise of the received code (Figure 13.5).

This method of measuring the carrier phase is referred to as integrated Doppler and is performed on each satellite. By integrating the Doppler frequency of the carrier $d\varphi/dt$, φ is obtained. This is the phase shift that changes with time, which is directly proportional to the change of range with time. Therefore, this phase plotted with time has the same slope as range plotted with time. The absolute range is not required, since the phase measurements are used only to smooth out the code solution. The low-noise carrier integrated Doppler plot (corresponding to range change with time) is better than the noisy code solution of the range, which is also

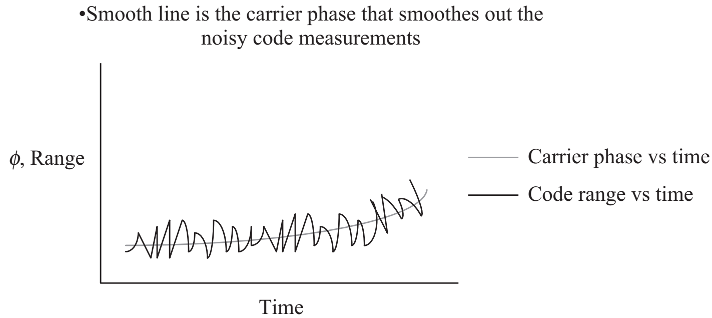

Figure 13.5 Carrier smoothed code using integrated Doppler

changing with time. The Doppler is a changing frequency, and integrating a frequency change produces a phase change, which is plotted on the same graph as the range change with time (Figure 13.5).

Another way to evaluate the integrated Doppler is to simply use the frequency times the period of integration:

$$\text{Integrated Doppler} = \int f\,dt = f \times t$$

For example, if the Doppler frequency is 1 kHz and the time of integration is from 0 to 0.5 s, then the equation becomes:

$$\text{Integrated Doppler} = \int_0^{0.5} 1{,}000\,dt = 1{,}000 \times t\big|_0^{0.5} = 1{,}000 \times 0.5 = 500\text{ s}$$

This is used to smooth out the code variations. For example, if the satellite is moving away from the receiver, the range (R) is positively increasing as the phase (φ) is negatively increasing. If the satellite is moving toward the receiver, the range is positively decreasing while the phase is positively increasing. Note that the Doppler is negative for a satellite going away, so the phase (φ) is negative yet the range is positive. A Kalman filter is often used to incorporate the integrated Doppler in the code solution. The Kalman filter is designed to achieve minimum variance and zero bias error. This means that the filter follows the mean of the time-varying signal with minimum variation to the mean. The Kalman filter uses the phase change to smooth the estimated range value of the solution as follows:

$$R_a = R_e - \text{Integrated Doppler}$$

where R_a is the actual range, R_e is the measured range.

A filter could be used to smooth the noise of the code, but it would be difficult to know where to set the cutoff frequency to filter out the changes since this would vary with each of the satellites being tracked.

13.9 Differential GPS

Differential GPS (DGPS) uses a differencing scheme to reduce the common error in two different receivers. For example, if the GPS receivers are receiving approximately the same errors due to ionosphere, troposphere, and SA, these errors can be subtracted out in the solution. This would increase the accuracy tremendously, since these errors are the main contributors to reduced accuracy. As noted earlier, this technique was used to virtually eliminate the problems with SA, which has now been turned off. The main contributor to errors presently is the ionosphere.

DGPS uses a ground-based station with a known position. The ground station calculates the differential pseudorange corrections, that is, how the pseudoranges are different from the surveyed position. These corrections are sent to other non-surveyed GPS receivers, such as an aircraft or ground vehicle. These pseudorange corrections are then applied to the aircraft's or ground vehicle's pseudoranges for the same satellites (Figure 13.6). The accuracy obtained by differential means can produce a position solution that is approximately 100 times more accurate than an autonomous C/A code receiver: approximately 1 to 3 m worst case.

13.10 DGPS time synchronization

The GPS time provided in the satellite navigation data is used to synchronize the receiver's time reference. SA can affect the time reference. DGPS helps to reduce SA effects on both time and orbital information. It is assumed that the time stamping of the aircraft and ground station occur at the same time even though

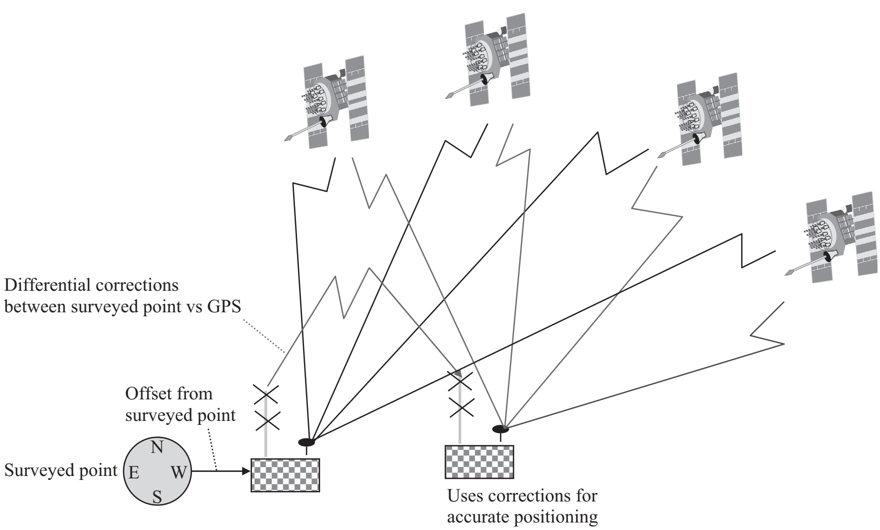

Figure 13.6 Differential GPS system for accurate positioning

absolute time is varied by SA. Currently, with SA turned off, the time synchronization is not affected by SA.

13.11 Relative GPS

Relative GPS is the determination of a relative vector by one receiver given its own satellite measurements and also the raw pseudorange measurements from the second receiver. The objective is to produce an accurate positioning solution between two systems relative to each other and not absolute positioning of each system.

For example, if an aircraft desires to land on an aircraft carrier using GPS, the aircraft carrier pseudoranges are uplinked to the aircraft. The aircraft calculates its relative position using its own measured pseudoranges with the pseudoranges that were sent by the aircraft carrier. The positioning errors that are in the pseudoranges from the aircraft carrier will be basically the same as the pseudoranges in the aircraft, so when the relative position solution is calculated, the errors are subtracted out. To enhance accuracy and integrity, the same set of satellites must be used so that the errors are common.

In a relative solution, atmospheric effects are assumed to be identical for both the ground, or aircraft carrier, and airborne units. Therefore, no tropospheric or ionospheric corrections are performed on the independent measurements. In a relative GPS system, the absolute accuracy depends on the accuracy of both the ground and airborne receivers. However, the absolute position does not need to be known, only the relative position with respect to each other.

Suppose that the GPS positioning solution for one GPS is off by 20 m toward the north. If the same satellites are used in the calculation of the GPS positioning solution for the other GPS, the assumption is that the other GPS would be off by 20 m toward the north. Therefore, both of their absolute positions would be off by 20 m toward the north. Since they are off by the same amount, their relative positions would be accurate since the errors that caused them to be off are close to the same and can be subtracted out (Figure 13.7).

The accuracy for a relative GPS is equivalent to the accuracy for a DGPS, or approximately 100 times more accurate than standard GPS—approximately 1 to 3 m worst case. Therefore, using relative GPS, what matters is not the absolute accuracy but only the relative accuracy between the aircraft and the aircraft carrier (Figure 13.8). Both relative GPS and DGPS use two different GPS receivers for accuracy. However, DGPS uses the difference from a surveyed point to calculate the errors needed to update the other GPS receiver, and the relative GPS uses the difference in the two GPS receivers to cancel the common errors.

13.12 Doppler

Doppler or range rate caused by the satellite moving toward or away from the GPS receiver needs to be determined so that it does not cause a problem in carrier phase tracking systems or when the carrier is used to smooth the code. Note that in a

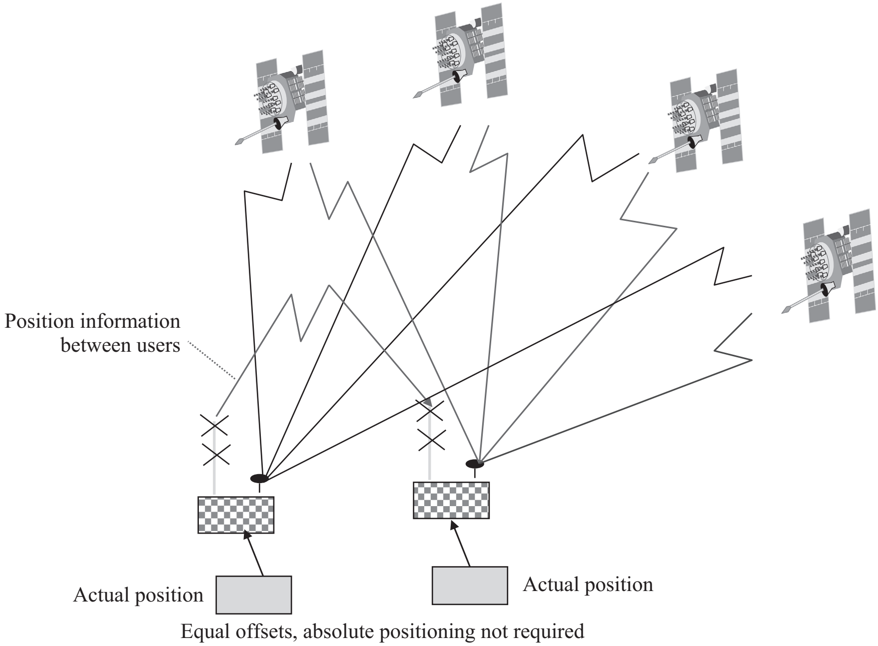

Figure 13.7 Relative GPS providing accurate relative positioning

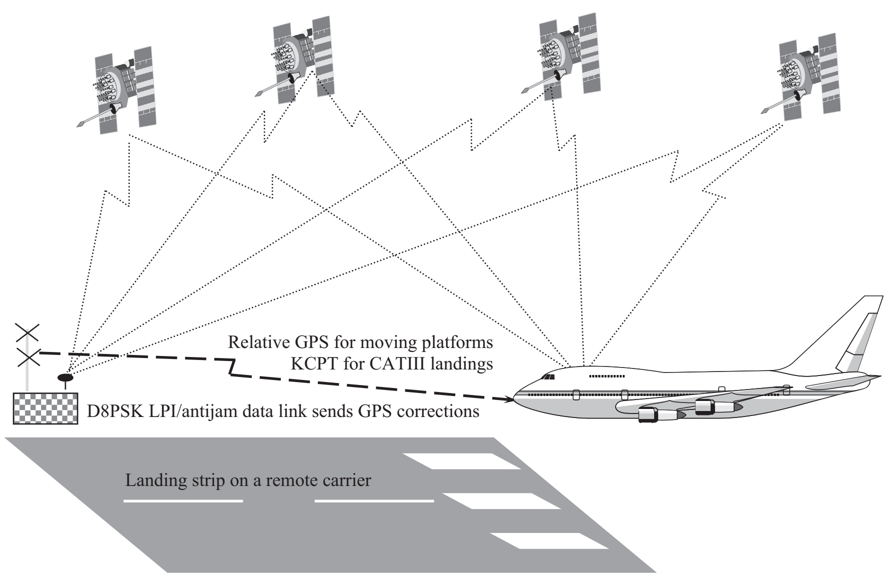

Figure 13.8 Relative GPS application for remote landing on an remote carrier

DGPS or relative GPS system, this Doppler is corrected in the process. The Doppler frequency range is about ±5 kHz for the worst case. Doppler causes the number of cycles to be off when using the carrier phase approach. If Doppler is known, then the number of cycles is known but not λ ambiguities inherent with carrier phase tracking.

13.13 Kinematic carrier phase tracking

Kinematic carrier phase tracking (KCPT) is used for very high-accuracy positioning, less than 10 cm. KCPT is most commonly used in survey applications but has recently been applied to high-accuracy commercial aircraft approach and landing systems. The KCPT solution uses the difference in phase of the carrier to determine range. The problem with the KCPT solution is the cycle ambiguity or wavelength ambiguity. The phase difference may be known, but the number of cycles for the range is not. But there are techniques to resolve the ambiguity of the number of cycles. If the receiver is moving, both the Doppler caused by the satellite and the Doppler due to receiver movement need to be considered.

One of the concerns with the KCPT solution after acquisition is what is known as cycle slip, especially when it is undetected. This is where the number of calculated cycles changes. If the cycle count slips, the range error is off by the number of cycles that have slipped. For L1, one cycle slip affects the pseudorange by approximately 0.6 ft or 19 cm. The system needs to detect cycle slip to adjust the cycle number, maintain accuracy and continuity of function, and ensure the integrity and safety of the application.

There are several ways to detect cycle slip and correct the number of cycles, one of which is using a differential autonomous integrity monitor, where the ground observables are compared with the airborne observables for each satellite. Another way involves performing multiple solutions in parallel using the same raw measurement data and then comparing them. And other real-time tests considering bias and noise can be performed. Regardless of the method used to detect cycle slips, the ambiguities must be resolved again using some a priori data from the previous resolution.

13.14 Double difference

The double difference is used to solve for the number of wavelengths to prevent cycle slip. The process involves taking the phase difference between satellites and subtracting the result from the baseline difference between antennas on the ground or air. Any two receivers can be used for the double difference. The main objective is to solve for N, the number of wavelengths:

$$\nabla\Delta = \Delta\varphi_1 - \Delta\varphi_2 = b(e_1 - e_2) + N\lambda$$

where b is the baseline vector and e_1, e_2 is the unit vectors to satellites.

$$\Delta\nabla = (\phi_1 - \phi_2)_{\text{Rec1}} - (\phi_1 - \phi_2)_{\text{Rec2}}$$
$$\Delta\nabla = \Delta\phi_1 - \Delta\phi_2 = b \cdot (e_1 - e_2) + N\lambda$$
$$\Delta\phi = \rho + \delta\rho + c(\delta t - \delta T) + \lambda N - d_{\text{ion}} + d_{\text{trop}} + \varepsilon\phi$$

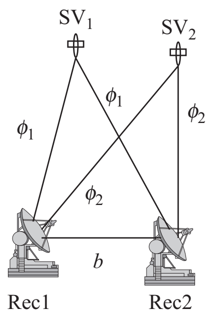

Figure 13.9 Double difference for kinematic carrier phase ambiguity resolution

Given the following absolute phase measurements:

$$\varphi = \rho + d\rho + c(dt - dT) + \lambda V - d_{\text{ion}} + d_{\text{trop}} + \varepsilon\varphi$$

where ρ is the geometric range, $d\rho$ is the orbital errors, dt is the satellite clock offset, dT is the receiver clock offset, d_{ion} is the ionosphere delay, d_{trop} is the troposphere delay, and ε is the noise.

There are two ways to achieve the double difference. One is to take the difference for one receiver and two satellites for first the carrier phase and then the code:

$$\varphi = \delta\rho + \delta d\rho + c\delta(dt - dT) + \lambda\delta V - \delta d_{\text{ion}} + \delta d_{\text{trop}} + \varepsilon\delta\varphi$$
$$\delta P = \delta\rho + \delta d\rho + c\delta(dt - dT) + \delta d_{\text{ion}} + \delta d_{\text{trop}} + \varepsilon\delta\varphi$$

The d_{ion} in the top equation is very small or zero in many cases if the ionospheric effects are the same for both the aircraft and the ground station. The second receiver does the same, and then the resultants are subtracted to achieve the double difference (Figure 13.9).

Another approach is to take the difference between the two receivers and one satellite and then between the two receivers and the next satellite. Finally, take the difference of the two resultants.

Note that the atmospheric losses are reduced or eliminated by receiver differences. For short antenna baselines, they are eliminated. For large baselines, they are reduced inversely proportional to the separation of the antennas.

13.15 Wide lane/narrow lane

Wide lane is used to reduce the cycle ambiguity inherent in the carrier phase tracking process. It uses the difference in the received frequencies L1 and L2

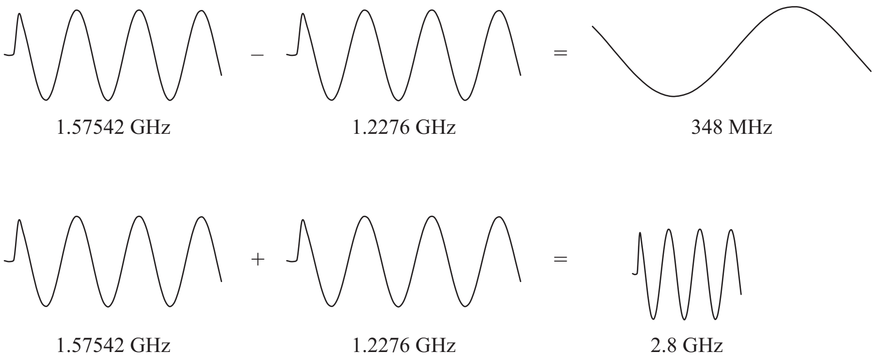

Figure 13.10 Wide lane and narrow lane for GPS waveforms

(L1 – L2), which produces a lower frequency (348 MHz) containing a larger wavelength, approximately 3.5 to 4.5 times greater (Figure 13.10). Therefore, the ambiguity search time is 3.5 to 4.5 times smaller. The lower frequency contains less cycle ambiguities because there are fewer cycles over the same distance. However, the disadvantage is that the accuracy is reduced.

Narrow lane uses the combination of L1 and L2 (L1 + L2), which provides more accuracy at the expense of more ambiguities to search over (Figure 13.10). A combination can be used: wide lane for ambiguity search and narrow lane for accuracy.

The frequencies and wavelengths are shown as follows:

L1 = 1.57542 GHz = 19 cm
L2 = 1.2276 GHz = 24 cm
L1 – L2 = 348 MHz = 86 cm
L1 + L2 = 2.8 GHz = 11 cm

13.16 Other satellite positioning systems

Two other satellite positioning systems use their own satellites in space to provide positioning. Many satellite receivers today are capable of receiving these additional satellite positioning systems for better coverage worldwide.

The GNSS developed by the Soviet Union, which is known as GLObal'naya NAvigatsionnaya Sputnikovaya Sistema (GLONASS), has been in operation since the mid-1990s. The system consists of 24 medium earth orbit satellites (MEOS): 21 are used for positioning, with 3 spares. However, the total number of operational satellites has varied over the years. The satellites are split into three inclined orbital planes separated by 120°, with an orbital time of 11 h and 15 min. The GLONASS consists of a low-accuracy C/A code and a high-accuracy P-code for military use. The low-accuracy code is similar to GPS stand-alone C/A code with SA turned on, less than 100 m. The high-accuracy P-code provides accuracy of about 10 m. The GLONASS operates in two frequency bands—1,602.5625–1,615.5 MHz

and 1,240–1,260 MHz—using 25 channels separated by 0.5625 MHz and frequency division multiplexing for separation of users.

Galileo In-Orbit Validation Element is a European satellite positioning system. The first satellites are in orbit, and the plan is to have a total of 30 satellites in orbit. This satellite system shares frequency bands with the United States and provides separation of users by code division multiplexing using different pseudonoise codes for each satellite.

13.17 Summary

GPS technology is being used for many applications including surveying, air traffic control and landing, position location for hikers, and mapping and location functions for automobiles. Originally developed for military applications, GPS has been adapted for commercial use using mainly C/A code receivers. Industry is now focusing more effort on using GPS signals in innovative ways that enhance the accuracy of measurements and therefore improve the availability and integrity of user services. Other positioning systems are discussed.

13.18 Problems

1. What is the null-to-null bandwidth of C/A code and P-code GPS signals?
2. What is the theoretical process gain of C/A code and P-code signals using a 50-Hz data rate?
3. What are the advantages of using C/A code over P-code?
4. What are the advantages of using P-code over C/A code?
5. What is the advantage of using the narrow correlator detection process?
6. Name at least two disadvantages of using the narrow correlator detection process.
7. Name the two harmful effects of SA on a GPS receiver.
8. What is carrier smoothing? Why is the code noisier than the carrier?
9. What is the main reason that DGPS is more accurate than standard GPS?
10. What is the main obstacle in providing a KCPT solution?
11. Why is the wide lane technique better for solving wavelength ambiguities?
12. How are multiple pseudonoise codes produced for GPS systems?
13. How are common errors reduced in GPS systems to obtain more accurate results?

Further reading

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Chapter 14

Introduction to radar and radar communications

RAdio Detection And Ranging (RADAR) is a method of using electromagnetic waves to determine the position of a target. Radar transmits a signal and receives and detects a portion of the signal that is reflected back to the radar. Radar uses this return signal or echo to measure the time it is transmitted to the time it is received to determine the range of the target. In addition, the returned signal can be received by the radar antenna to determine the angle it received. Therefore, radar can determine the range and direction, velocity, and identifying characteristics of targets by monitoring the reflected signals coming back to the radar.

Radar was developed for military purposes during WWII and both the British and US Militaries used radar to locate ships and airplanes. Radar operators detected false alarms due to weather returns and didn't realize how sensitive radar is to rain and moisture. This phenomenon later became a useful tool in weather radars. A short history of radar with key milestones began with the development of electromagnetic light theory by English physicist James Clerk Maxwell as follows:

- 1865—Developed electromagnetic light theory by English physicist James Clerk Maxwell
- 1886—Discovers electromagnetic waves and Maxwell's theory by German physicist Heinrich Rudolf Hertz
- 1904—First practical radar test for distance which measured time of electromagnetic waves to a metal object (ship) and back. German Christian Hülsmeyer receives patent for the "Telemobiloskop" or Fernbewegungseher
- 1917—Superheterodyne receiver invented by French engineer Lucien Lévy
- 1921—Magnetron invented by American physicist Albert Wallace Hull
- 1922—Located a wooden ship by American electrical engineers Albert H. Taylor and Leo C. Young of the Naval Research Laboratory
- 1930—Locates an aircraft by Lawrence A. Hyland of the Naval Research Laboratory
- 1931—Ship equipped with radar with a parabolic dish antenna with horn radiator
- 1936—Klystron developed by George F. Metcalf and William C. Hahn, General Electric
- 1940—Radar equipment developed by USA, Russia, Germany, France, and Japan

14.1 Radar applications

Both the military and commercial entities have multiple applications using radar. The military use radar for search and detection of targets, target tracking, missile guidance, and fire control including acquisition and track. It is also used for intercepting aircraft, ground and battle field surveillance, air and space mapping, and many others. Submarines and subchasers have used radar for location of submarines and other platforms in their military operations.

The commercial world use radar for weather, navigation, and air traffic control. Radar is used on the road and highways for detection of vehicles going over the speed limit. In addition, they are used for search, acquisition, and track of commercial aircraft. Radar is useful in biological research, bird, and insect surveillance and tracking and is used in the medical field for diagnosis, organ movements, water condensation in the lungs, monitor heart rate, and pulmonary motion.

Also included in the medical field are range for operations, remote sensor of heart and respiration rates without electrodes, and for patient movement to detect falls at home. Miniature radars are also used for seeing aids, early warning collision detection, and situational awareness.

14.2 Two basic radar types

There are two basic types of radars that are used extensively today. One is continuous wave radar which sends out a continuous wave signal and receives a reflected signal off of the moving targets which produces a Doppler frequency. This type of radar requires two receive antennas with one of the antennas pointed at the transmit antenna and one is pointed at the target. The one pointed at the transmit antenna is used to cancel the transmitted signal which leaves the Doppler information for detection. The continuous antenna generally does not determine range or altitude information, it is mainly used to determine movement and change. This type of radar has a high signal-to-noise ratio (SNR) so it is more difficult to jam, however, because it is a continuous waveform it can be easily deceived. It is simpler to operate because precise timing is not required.

The other type of radar is pulse radar which transmits a pulse stream with a low duty cycle and monitors the received reflections from targets during the time off or dead time between pulses. This type of radar uses a single antenna for both transmit and receive since it is not receiving during transmission, and by monitoring the delays of the reflection it can determine range and altitude. It is susceptible to jamming because the jammer can determine the duty cycle and can jam only during the time the pulse is on.

Most radar systems are pulsed radars. This avoids the problem of a sensitive receiver simultaneously operating with a high power transmitter. The pulse radar transmits a low duty cycle, short duration high-power RF pulses. The time of the transmit pulse is compared to the time of arrival of the received reflected pulse to determine the range of the target.

14.3 Basic pulse radar operation

The basic pulse radar uses a directional antenna pointed toward a desired target. The radar sends out the radar pulse toward the target, and then receives the reflection or echo from the target (Figure 14.1). The radar consists of a transmitter for generating the output pulse and a low noise receiver to detect the time when the radar pulse returns. The receiver requires a low-noise figure in order to detect low-level signals. The transmitted pulse travels through a transmit channel and reflects off the target which has a reflective radar surface. The returned signal is processed by the low noise receiver (Figure 14.2).

The pulse radar transmits a high-level pulse with a low duty cycle so the average transmit power is low. The pulse radar requires time synchronization to determine range to the target. Pulses are repeated at the Pulse Repetition Frequency (PRF) (Figure 14.3). This is the basic operation of pulsed radar. The PRF is the number of pulses per second and the Pulse Repetition Interval (PRI) is the time between the start of the pulses. Another term used in radar is the Pulse Repetition Time which is the same as PRI.

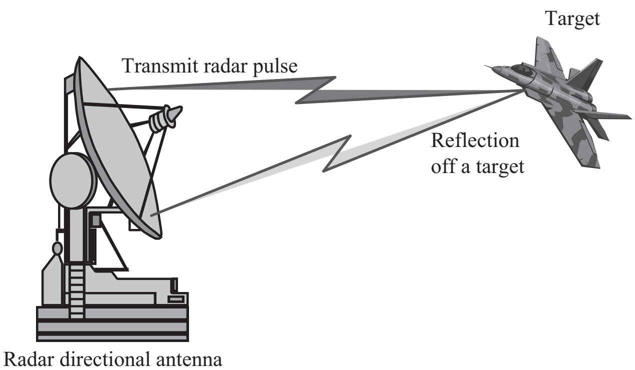


Figure 14.1 Basic radar operation

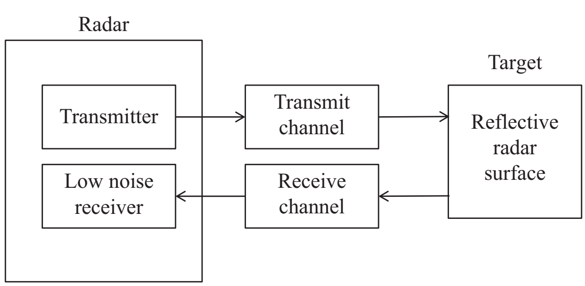


Figure 14.2 Radar functional block diagram

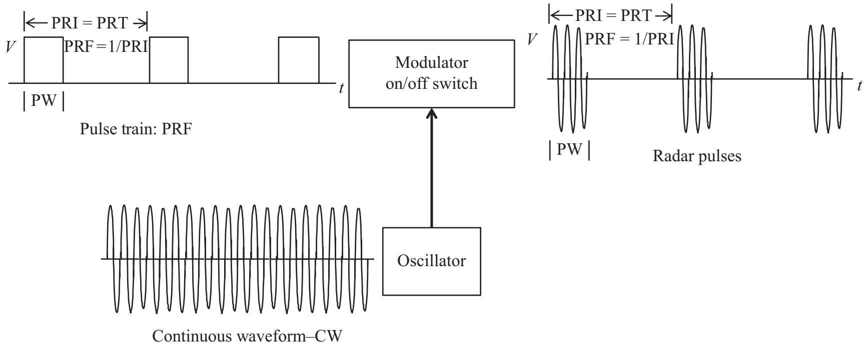


Figure 14.3 Basic operation to generate standard radar pulses

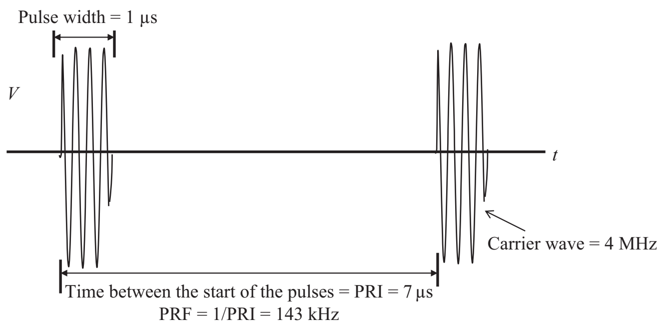


Figure 14.4 On/off keying of a CW waveform produces the radar pulses

14.3.1 Pulse radar modulation

The modulation is referred to as on/off keying. Basically, radar pulse turns on/off a carrier frequency or local oscillator (LO). This creates 100% Amplitude Modulation (AM) signal with the modulation being a pulse with a low duty cycle (Figure 14.4). The Pulse Width (PW) is how long the carrier is on or the amount of time that the radar is transmitting, the PRI is the interval or time from the start of one pulse to the start of the next pulse, and the PRF is the frequency of the modulating pulse waveform which is equal to 1/PRI (Figure 14.4). The example in the figure shows a carrier wave equal to 4 MHz, a PW of 1us, and a PRF of 143 kHz. The duty cycle is the percentage that the pulse is on compared to the total time which is equal to the PW/PRI. The calculations are shown below:

Given: PW = 1 μs, PRI = 7 μs
PRF = 1/PRI = 1/7 μs = 143 kHz
4 carrier cycles = 1 μs; 1 cycle = 1 μs/4 = 250 ns; f_o = 1/250 ns = 4 MHz
Duty cycle = PW/PRI = (1 μs/7 μs) × 100 = 14.3%

where PW is the pulse width, PRI is the pulse repetition interval, PRF is the pulse repetition frequency, and f_o is the carrier frequency.

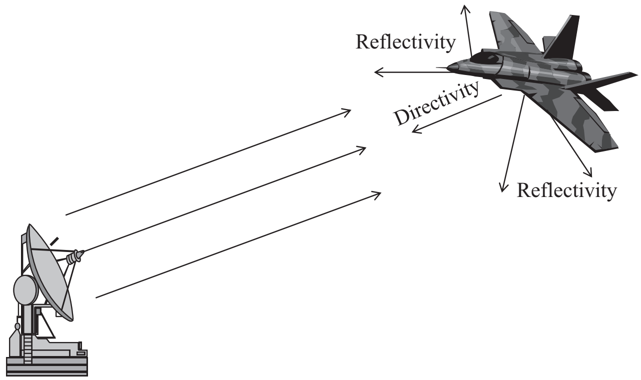


Figure 14.5 Radar cross section (RCS) determines the level of reflected signal

14.3.2 Radar cross section

All targets scatter electromagnetic energy and some of the energy is scattered back toward the radar. The amount of energy that is scattered back to the radar depends on its radar cross section (RCS). Each target has its own RCS, represented by σ, and is equal to:

$$\sigma = \text{Projected cross section} \times \text{reflectivity} \times \text{directivity}$$

where projected cross section is the shape and size of the cross section, reflectivity is the ability to reflect the RF signal, and directivity is the ability to focus energy back to the radar—similar to antenna gain.

Reflectivity and directivity are similar to the gain of an antenna and the RCS is a gain factor in the radar budget analysis. The target radar cross-sectional area depends on the size and the ability of a target to reflect radar energy which equals the target's cross-sectional area theoretically (Figure 14.5). It also depends on the material type of the reflecting surface, the direction of the illuminating radar, and the transmitted frequency. However, not all reflected energy is returned back to the radar since the energy is scattered or distributed in all directions with some energy being absorbed.

The shape of the target is a key factor in determining the RCS for the reflector (Figure 14.6). The corner reflector has the highest RCS with the flat panel second highest, and the sphere the lowest RCS. However, there are many shapes other than these standard three and it becomes very difficult to estimate. For example, an aircraft's physical geometry and exterior features makes it difficult to determine the RCS, and if it is moving, the shape is actually changing. So often times the best approach is to measure the RCS for a given surface and keep these measurements as a database for evaluating the RCS for a target. There are also software programs that incorporate different RCSs for different platforms to approximate the RCS.

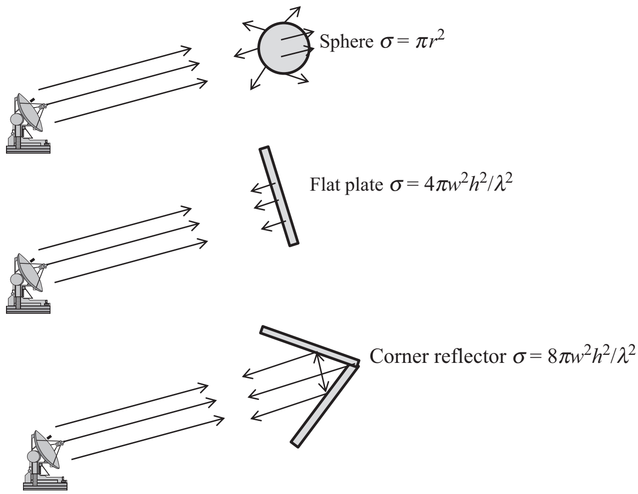


Figure 14.6 RCS patterns are dependent on the shape of the object

14.3.3 Radar path budget

This is the basic path of the radar pulse, and the radar path budget tracks the pulse signal level and the noise level through this path. The radar path budget begins with the power out of the transmitter power amplifier (PA), includes the losses in the transmitter to the antenna, and incorporates the gain of the antenna. This is referred to as the Effective Isotropic Radiated Power (EIRP). The Effective Radiated Power is referenced to a Dipole antenna instead of the Isotropic antenna so it will be equal to the EIRP minus the gain of a Dipole antenna (theoretically 2.14 dB). The ratio of the signal-to-noise (S/N) is evaluated at the receiver with the required SNR determining the ability to reliably detect the return signal. The radar budget uses the required S/N and makes adjustments to the total path to ensure that the actual S/N meets the requirement. This includes the EIRP out of the transmitter, the transmitter hardware losses, the transmitter channel losses, the target reflectivity, the receiver channel losses, and the receiver hardware losses. The signal is received with the required S/N level at the detector. This radar budget is used to solve trade-offs between all of the elements in the radar path, which includes size, cost, range, and others in addition to allocating the power and noise through the path. This radar budget is used with all available return signals.

In summary, the basic radar operation is as follows:

1. Radar transmits a high-level signal
2. Uses high PAs
3. Uses high gain directional antennas
4. Directional antenna radiates electromagnetic pulses toward the target
5. Pulses generally travel line-of-site at the speed of light
6. Pulses reflect off the target back to the radar at the same speed

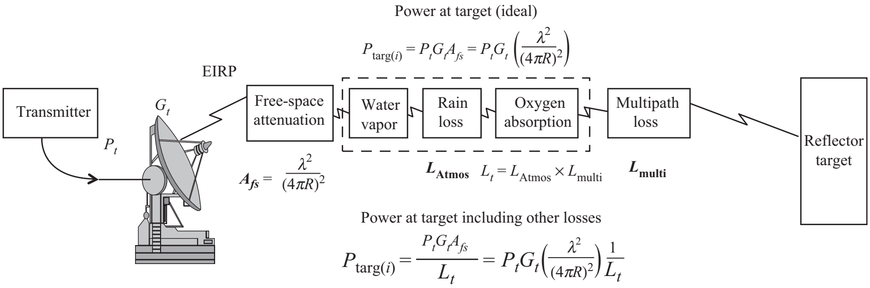


Figure 14.7 Radar transmitter power to target

7. Receiver captures the return (echo) and processes the signal
8. Low noise receiver used to optimize the detection of the echo
9. Receiver calculates the Slant Range and Angle of Arrival (AoA)
10. High isolation protects the receiver from the high power transmitter
11. Uses a duplexer, diplexer, or transmit/receive (T/R) switch to protect receiver damage

14.3.3.1 The transmit path

The transmitter path from the radar transmitter to the target is shown in Figure 14.7. The transmitter sends power to the directional antenna to generate the EIRP. The biggest loss to the signal is called freespace loss. The freespace loss equation is given in several different forms depending on how it is used and whether it is less than one and multiplies the signal or it is greater than one and divides the signal. In addition, it can also be expressed in dB and if it is negative, it is added to the signal and if it is positive it is subtracted from the signal. The following are the equations forms of freespace loss:

$A_{fs} = (\lambda/(4\pi R))^2$ will be less than 1 and multiplied
$A_{fs} = ((4\pi R)/\lambda)^2$ will be greater than 1 and divided
$A_{fs} = 10\log\,(\lambda/(4\pi R))^2 = 20\log\,\lambda/(4\pi R) =$ will be a negative number and added
$A_{fs} = 10\log\,((4\pi R)/\lambda)^2 = 20\log\,(4\pi R)/\lambda =$ will be a positive number and subtracted

where A_{fs} is the freespace loss, λ is the wavelength, and R is the slant range.

It is important to determine which of these forms the freespace loss is given or used.

The following are examples of using the different forms of freespace loss:

Given: $P_t = 100$ W $= 50$ dBm, $\lambda = 0.125$, $R = 1{,}000$ m
$A_{fs} = (\lambda/(4\pi R))^2 = (0.125/(4\pi 1{,}000))^2 = 98.9 \times 10^{-12}$: $P_r = 100$ W $\times\ 98.9 \times 10^{-12} = 9.89 \times 10^{-9}$
$A_{fs} = ((4\pi R)/\lambda)^2 = ((4\pi 1{,}000)/0.125)^2 = 1.01065 \times 10^{10}$: $P_r = 100$ W$/(1.01065 \times 10^{10}) = 9.89 \times 10^{-9}$

$A_{fs} = 20\log\ \lambda/(4\pi R) = 20\log\ 0.125/(4\pi 1{,}000) = -100$ dB: $P_r = 50$ dBm + $(-100$ dB$) = -50$ dBm

$A_{fs} = 20\log\ (4\pi R)/\lambda = 20\log\ (4\pi 1{,}000)/0.125 = 100$ dB: $P_r = 50$ dBm − (100 dB) = −50 dBm

The power to the target is therefore:

$$P_{\text{target}} = P_t G_t A_{fs} = P_t G_t \frac{\lambda^2}{(4\pi R)^2}$$

where P_{target} is the power at the target, P_t is the radar transmitted power, G_t is the transmitter antenna gain, A_{fs} is the freespace loss, λ is the wavelength, and R is the slant range.

The other losses that need to be included are the atmospheric losses due to rain, vapor, oxygen absorption, and loss due to multipath (Figure 14.7).

$$P_{\text{target}} = \frac{P_t G_t A_{fs}}{L_t} = \frac{P_t G_t}{L_t} \frac{\lambda^2}{(4\pi R)^2}$$

where P_{target} = power at the target, P_t = radar transmitted power, G_t = transmitter antenna gain, A_{fs} = freespace loss, λ = wavelength, R = slant range, and $L_t = L_{\text{Atmos}} + L_{\text{Multi}}$ = total losses, where L_{Atmos} = atmospheric losses and L_{Multi} = multipath losses.

14.3.3.2 The receive path

Radar receives power from the target (Figure 14.8). The amount of power received is based on the transmitted power to the target, gain of the target due to the RCS and the losses associated with the receive path, and other losses in the hardware in the receiver. The freespace loss is the major contributor to the overall losses in the channel. There is further gain due to the radar receiving antenna which increases

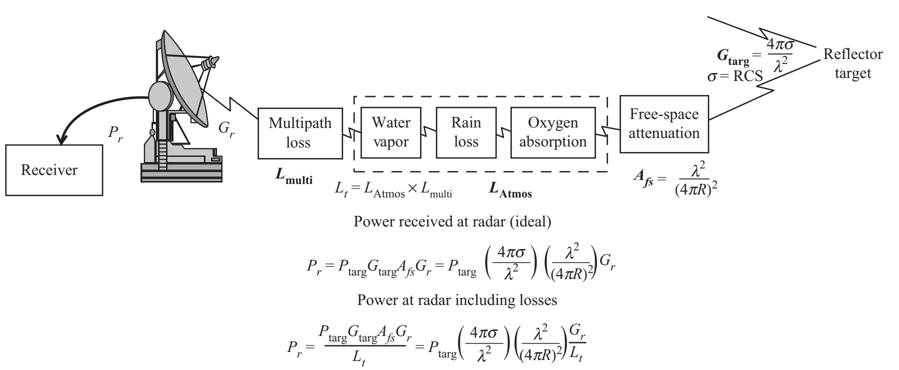


Figure 14.8 Radar received power from target

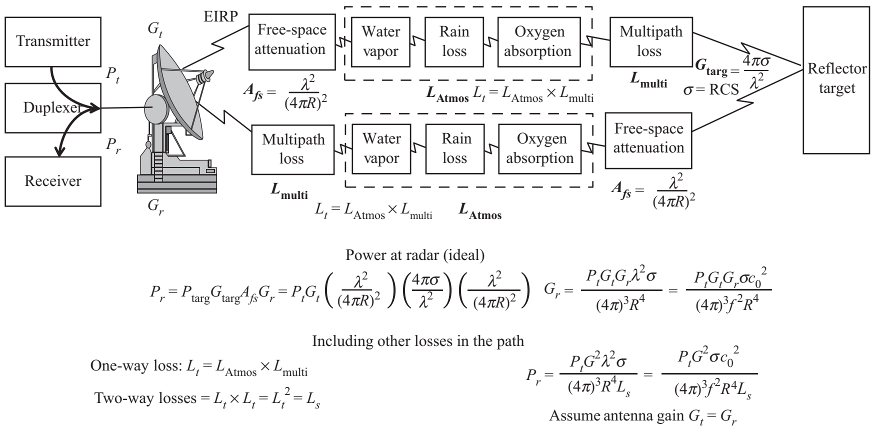


Figure 14.9 Radar antenna gain and channel losses included for total radar path

the power in the receiver (Figure 14.8). The equation for the received power using only the freespace loss is:

$$P_r = P_{\text{targ}} G_{\text{targ}} A_{\text{fs}} G_r = P_{\text{targ}} \frac{4\pi\sigma}{\lambda^2} \frac{\lambda^2}{(4\pi R)^2} G_r$$

where P_r is the radar received power, P_{target} is the power at the target, G_{target} is the gain of the target $= 4\pi\sigma/\lambda^2$, σ is the RCS, λ is the wavelength, A_{fs} is the freespace loss, G_r is the receiver antenna gain, and R is the slant range.

The other losses that need to be included are the atmospheric losses due to rain, vapor, oxygen absorption, and loss due to multipath (Figure 14.8).

$$P_r = \frac{P_{\text{targ}} G_{\text{targ}} A_{fs} G_r}{L_t} = P_{\text{targ}} \frac{4\pi\sigma}{\lambda^2} \frac{\lambda^2}{(4\pi R)^2} \frac{G_r}{L_t}$$

$P_r =$ radar received power, $P_{\text{target}} =$ power at the target, $G_{\text{target}} =$ gain of the target $= 4\pi\sigma/\lambda^2$, $\sigma =$ RCS, $\lambda =$ wavelength, $A_{fs} =$ freespace loss, $G_r =$ receiver antenna gain, $R =$ slant range, and $L_t = L_{\text{Atmos}} + L_{\text{Multi}} =$ total losses, where $L_{\text{Atmos}} =$ atmospheric losses and $L_{\text{Multi}} =$ multipath losses.

14.3.3.3 Combined total radar path

The power received at the radar substituting the power received at the target provides the total radar equation (Figure 14.9). The equation for the radar received power without the other losses is:

$$P_r = P_{\text{targ}} G_{\text{targ}} A_{fs} G_r = P_t G_t \frac{\lambda^2}{(4\pi R)^2} \frac{4\pi\sigma}{\lambda^2} \frac{\lambda^2}{(4\pi R)^2} G_r = \frac{P_t G_t G_r \lambda^2 \sigma}{(4\pi)^3 R^4}$$

$$= \frac{P_t G_t G_r \sigma {c_0}^2}{(4\pi)^3 f^2 R^4}$$

where λ = wavelength, $\lambda = f \times c_o$, f = carrier frequency, and c_o = speed of light = 3×10^8.

When including the losses in the final equation, both the transmit path and the receive path need to be accounted for with respect to noise. Therefore, there are two freespace losses, one for the radar transmitter to target path and one for the target to radar receiver path. There are also two path losses for both atmospheric loss and the multipath loss (Figure 14.9).

$$P_r = \frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 R^4 L_s} = \frac{P_t G^2 \sigma c_0{}^2}{(4\pi)^3 f^2 R^4 L_s}$$

where P_r = radar received power, P_t = power of the transmitter, G = gain of the antenna, λ = wavelength, $\lambda = f \times c_o$, f = carrier frequency, c_o = speed of light = 3×10^8, σ = RCS, R = slant range, and L_s = 2-way losses = $L_t \times L_t = L_t^2$, where $L_t = L_{\text{Atmos}} + L_{\text{Multi}}$ = total losses, where L_{Atmos} = atmospheric losses and L_{Multi} = multipath losses.

This is the standard radar equation that is used for radar applications.

There are other losses that are not specifically shown that need to be included. They are: the T/R switch or circulator loss, cable losses, antenna losses such as polarization, mispointing, radome and hardware implementation loss, and possibly others. Some of these losses will be in one of the paths or both and need to be included in the final budget.

An example using the standard radar equation to determine the power level in the receiver as follows (Figure 14.10).

Given: f = 2.4 GHz, $\lambda = 0.125$, P_t = 100 W, R = 1,000 m, $G_t = G_r$ = 1,000, $L_s = 10$, $\sigma = 140$ m^2

$P_r = 100(1{,}000)^2(0.125)^2(140)/(4\pi)^3(1{,}000)^4(10) = 1.10235 \times 10^{-8}$ W = 1.0235×10^{-5} mW

$P_{\text{rdbm}} = 10\log\,(1.10235 \times 10^{-5}) = -49.6$ dBm

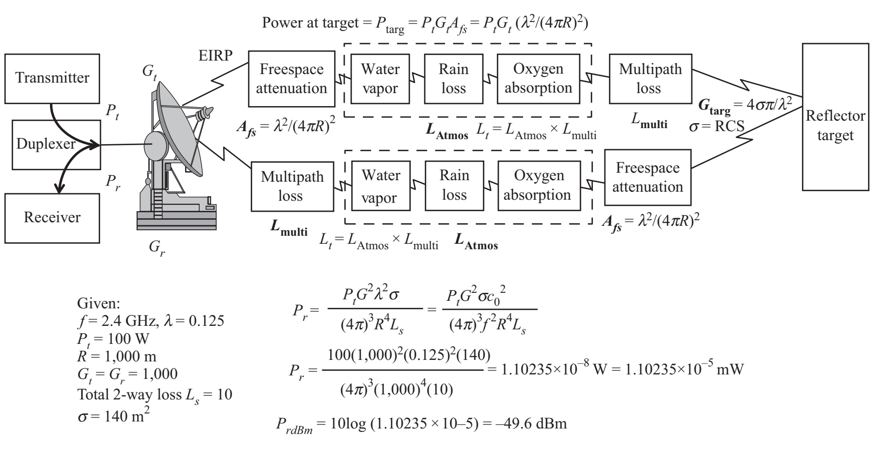


Figure 14.10 Radar example using the radar equations

In order to simplify the calculations, an alternative approach is to calculate the solution in dB as follows (Figure 14.11).

Given: $f = 2.4$ GHz, $\lambda = 0.125$, $P_t = 50$ dBm, $R = 1{,}000$ m, $G_t = G_r = 30$ dB, $L_s = 10$ dB, $\sigma = 140\ \text{m}^2$
$A_{fs\text{dB}} = 20\log\,[(0.125)/(4\pi 1{,}000)] = -100.05$ dB
$G_{\text{targ}} = 10\log\,(4\pi \times 140/0.125^2) = 50.5$ dB
$P_{r\text{dBm}} = 50\ \text{dBm} + 2 \times 30\ \text{dB} + 2 \times -100.05\ \text{dB} + 50.5\ \text{dB} - 10\ \text{dB} = -49.6$ dBm

An alternate equation is sometimes used which includes a pattern propagation factor as follows:

$$P_r = \frac{P_t G_t A_{er} \sigma F^4}{(4\pi)^2 R^4 L_s}$$

where P_r = radar received power, P_t = power of the transmitter, G_t = gain of the transmit antenna, A_{er} = radar receiver antenna effective aperture area ($A_r \times n$), Ar = area of the antenna, n = efficiency factor of the aperture, F = pattern propagation factor, R = slant range, and L_s = 2-way losses = $L_t \times L_t = L_t^2$, where $L_t = L_{\text{Atmos}} + L_{\text{Multi}}$ = total losses, where L_{Atmos} = atmospheric losses and L_{Multi} = multipath losses.

Also, the S/N equation is sometimes used as follows:

$$\text{SNR} = \frac{P_t G^2 \lambda^2 \sigma t_{\text{pulse}}}{(4\pi)^3 R^4 k T_s L_s}$$

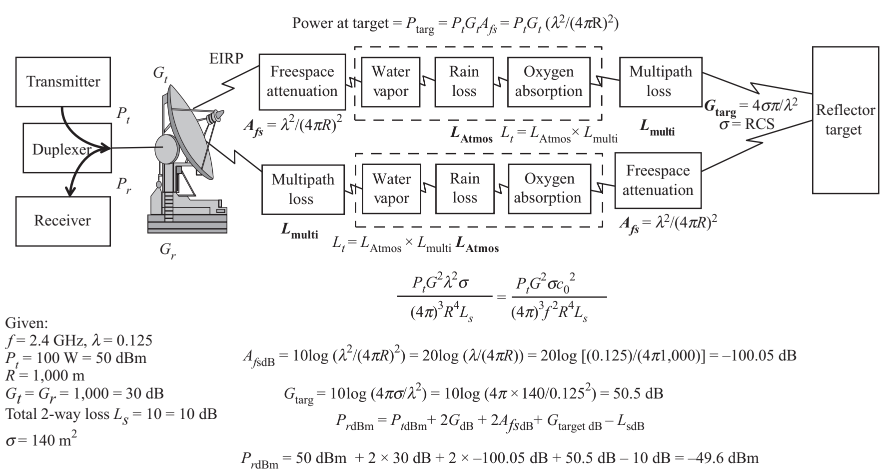


Figure 14.11 Radar example using the radar equations in dB

where SNR = signal-to-noise ratio, P_t = power of the transmitter, G = gain of the antenna, λ = wavelength, σ = RCS, t_{pulse} = duration of received pulse, R = slant range, k = Boltzmann's constant, T_s = system noise temp, and L_s = 2-way losses.

Noise $= k \times T_s$
Noise $= k \times T_s \times B \times$ NF if bandwidth and noise figure are specified.

14.3.4 Range determination

Radar range calculation uses the time delay which is measured from the time that the radar sends out the pulse to the time that it is reflected back and detected by the radar receiver. The radar compares the time that the pulse was transmitted to the time it is received. This time is the two-way time and is divided by two to provide the one-way time which is used to calculate the distance or range to the target.

14.3.4.1 Sound wave example

An example using sound waves instead of electromagnetic waves can be used to demonstrate the range calculation. If a person shouts in the direction of an object such as a mountain cliff, the sound is reflected off of the cliff back to the person (Figure 14.12). This reflection is called an echo. By measuring the time it took from when the person shouted and when the person hears the echo can be used to calculate the distance to the mountain cliff (Figure 14.12). The total round-trip time would be divided in half to calculate the one-way distance to the mountain. The following is an example on how to measure the distance using sound:

Given: Speed of sound = 1,125 ft/s
Measured time between the shout and hearing the reflected shout = 5 s
Distance = 5 s/2 × 1,125 ft/s = 2,813 ft

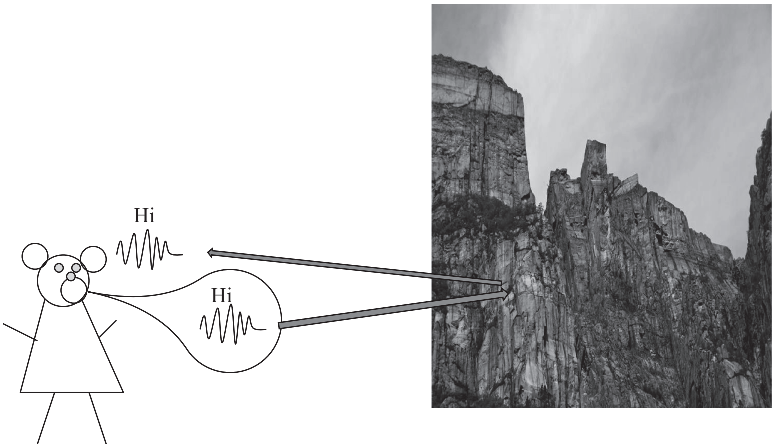


Figure 14.12 Sound echoes that reflect of canyon walls can determine the distance

Note that the most effective range measurement requires a short shout to prevent it from interfering with the echo. This is more critical for shorter distances.

Another example that is only one direction and not relying on a reflection is watching a lightning bolt and then listening to the thunder to determine the range of the lightning bolt. Since the speed of light is so much faster than sound, there is very little error. The one-way time between when the lightning bolt is visible to when the sound of the thunder is audible is used to calculate the range to the lightning bolt. For example:

Given: Speed of sound = 1,125 ft/s
Time between a visual of the lightning bolt and the audible of the thunder = 5 s
Distance = 5 s × 1,125 ft/s = 5,625 ft or just over a mile (1 mi = 5,280 ft)

Since the lightning example only uses one direction, the range calculation is not divided by two.

14.3.4.2 Radar range calculations

Radar uses electromagnetic energy pulses. The pulse travels at the speed of light c_0 which is 300 million meters/second, 984 million feet/second, or 186,411 mi/s. These pulses reflect off a surface and an echo is returned back to the radar receiver. The total two-way time is divided by 2 in order to calculate the one-way range as follows:

$$R_{\text{slant}} = t_{\text{measured}}/2 \times c_0$$

where R_{slant} is the slant range or the line-of-sight distance from radar to target, t_{measured} is the is the two-way measured time, and c_0 is the speed of light = 3×10^8 m/s.

An example of how to calculate the radar range is shown below:

Given: $t_{\text{measured}} = 1$ ms, $c_0 = 3 \times 10^8$ m/s
Calculate slant range = $R_{\text{slant}} = (1\text{ ms} \times 3 \times 10^8\text{ m/s})/2 = 150$ km

The basic equation includes refraction caused by transmitter wavelength, barometric pressure, air temperature, and atmospheric humidity.

The radar range equation is generated by using the basic radar equation below and solving for $R_{\max}$ using $S_{\min}$:

$$P_r = S_{\min} = \frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 {R_{\max}}^4 L_s}$$

where $P_r = S_{\min}$ is the radar minimum received power, P_t is the radar transmitted power, $G_t = G_r = G^2$ is the antenna gains, λ is the wavelength, $R_{\max}$ is the maximum slant range, σ is the RCS, L_t is the one-way losses = $L_{\text{Atmos}} + L_{\text{Multi}}$, $L_s =$ two-way losses = $L_t \times L_t = L_t^2$.

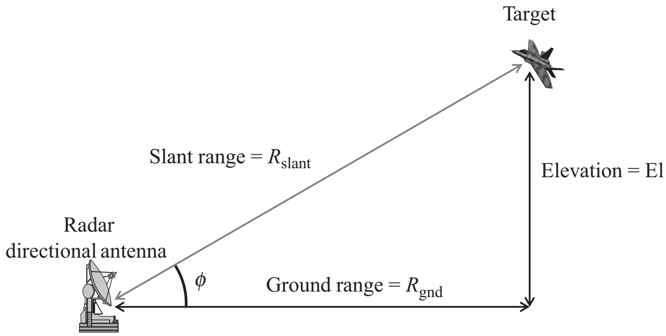


Figure 14.13 Slant range is the direct path to the target

Radar range equation (solving for $R_{\max}$ range for minimum signal $S_{\min}$):

$$R_{\max}{}^4 = \frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 S_{\min}{}^4 L_s}$$

$$R_{\max} = \left[\frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 S_{\min}{}^4 L_s}\right]^{1/4}$$

Note that double the range requires 16 times more transmit power P_t. The radar detection range is the maximum range at which a target has a high probability of being detected by the radar. The radar energy to the target drops proportional to range squared and the reflected energy to the radar drops by a factor of range squared. Therefore, the received power drops proportional to the fourth power of the range. Therefore, a radar requires very large dynamic ranges in the receive signal process. It also is required to detect very small signals in the presence of large interfering signals.

14.3.4.3 Slant range

The slant range is the direct path to the target. The ground range, if the target is in the air, is the horizontal distance from radar to target. The slant range can also be calculated using the ground range and elevation as follows (Figure 14.13). The slant range can also be calculated by using the elevation and the angle between the R_{gnd} and R_{slant}.

$$R_{\text{slant}}{}^2 = R_{\text{gnd}}{}^2 + \text{El}^2$$

$$R_{\text{slant}} = \left(R_{\text{gnd}}{}^2 + \text{El}^2\right)^{1/2}$$

$$R_{\text{slant}} = \text{El}/\sin\,\phi$$

where R_{slant} is the slant range, R_{gnd} is the ground range, El is the elevation, and ϕ is the angle between the R_{gnd} and R_{slant}.

Height is the distance above the ground or earth's surface and altitude is the distance above mean sea level (Figure 14.14). These terms are sometimes used

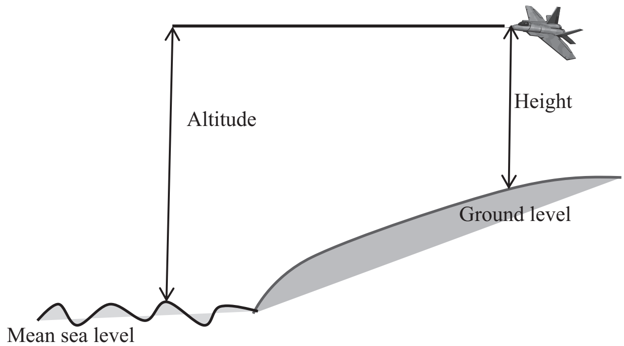


Figure 14.14 Height and altitude calculations

incorrectly with altitude replacing height. Calculating altitude requires Earth Center Earth Fixed (Figure 14.15). The elevation angle (ε) is from the horizontal baseline to the slant range vector, or the angle between the horizontal plane and line-of-sight to the target. $\varepsilon = 0°$ on the horizon, a positive angle above the horizon and a negative angle below the horizon The height is calculated as follows:

$$H = R \times \sin(\varepsilon) + R^2/(2r_e)$$

where H is the height, R is the slant range, ε is the measured elevation angle, r_e is the earth's equivalent radius (6,370 km).

14.3.4.4 Range ambiguity

Range ambiguity is caused by strong targets at a range in excess of the pulse repetition indicator or time. The pulse returns from the first transmitted pulse appear after the second pulse is transmitted so that the return is ambiguous of whether it was a return from the first pulse sent or the second pulse sent (Figure 14.16). This causes the range calculation to be close instead of farther away since the basic radar does not know which pulse is being returned. However, there are other advanced techniques used by certain radars in order to distinguish the different radar pulses. Large pulse amplitude and a higher PRF amplify the ambiguity problem. The maximum unambiguous range for given radar system can be determined by using the formula:

$$R_{\max} = (\mathrm{PRI} - T) \times C_0/2$$

where $R_{\max}$ is the maximum unambiguous range, PRI is the pulse repetition indicator, T is the pulse width time, and C_0 is the speed of light.

An example is as follows:

Given: PRI = 1 ms, $T = 1$ μs
Maximum unambiguous range = (1 ms–1 μs) × 3 × 10^8/2 = 149.9 km

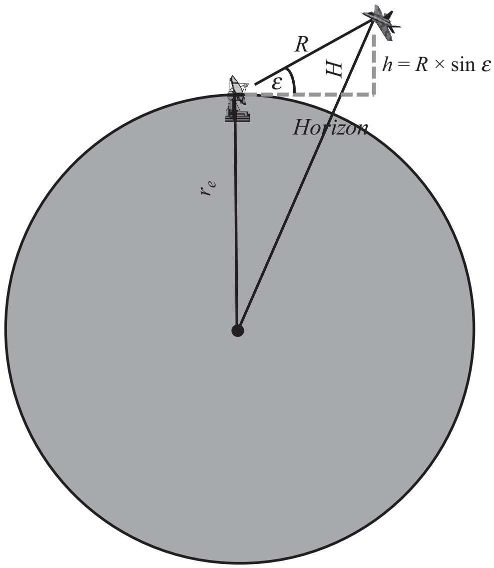


Figure 14.15 Earth center earth fixed altitude calculations

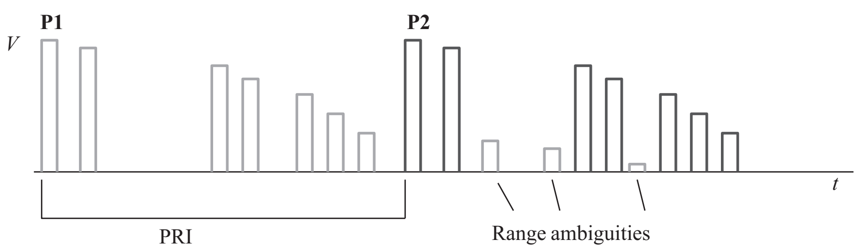


Figure 14.16 Range ambiguity with overlapping returns

In addition, this range ambiguity also can cause interference by the return of the first pulse to the returns of the second pulse. The design needs to ensure that the returns that come from the first pulse do not overlap the returns of the second pulse which causes distortion in the returns of the second pulse (Figure 14.17). Therefore, the PRF is set low enough so that the returns from the first pulse are sufficiently low in amplitude to prevent interference with the second pulse returns.

One way to help mitigate range ambiguity and interference is by decreasing the PRF which reduces the range ambiguity because the longer the time delays, the higher freespace loss which creates a smaller return.

Another way to help mitigate range ambiguity and interference is by transmitting different pulses with different modulation at each PRF interval. This

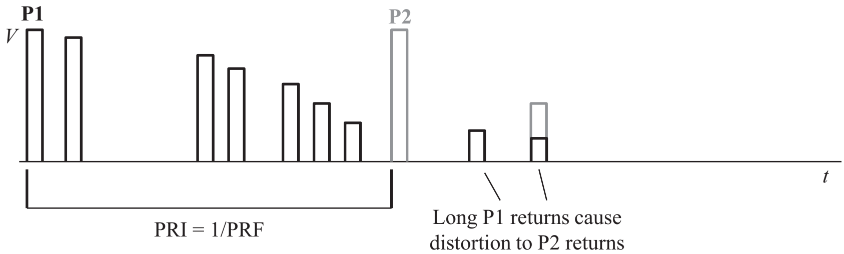


Figure 14.17 Pulse distortion caused by long returns after the next pulse

increases the receiver complexity and requires multiple matched filter correlators at each range bin and at each azimuth and elevation and it also increases the rate of the Digital Signal Processor (DSP) required for each separate transmit pulse and matched filter correlator pair.

Another technique to address the problem is to vary the PRF, depending on the radar's operational mode. This requires changing the system parameters but is used most often to mitigate range ambiguity.

Generally, the maximum range that the radar can detect is determined by the PRF. The PRF determines the radar's maximum detection range since long returns from the first pulse can cause distortion to the returns of the second pulse sent.

14.3.4.5 Minimum range

The minimum range that radar can detect is based on the transmitter PW. The PW establishes the minimum range since a radar return cannot come back during the transmitted pulse width or it will be distorted in the receiver (Figure 14.18). The returns R1 and R2 are distorted by the transmitted pulse P1. The return R3 is the minimal detectable range and is not distorted by the transmitted pulse since it is not present. Therefore, the minimum detectable range is equal to:

$$R_{\min} = \left(T + T_{\text{recovery}}\right) \times c_0/2$$

where T is the pulse width, T_{recovery} is the time for pulse to recover, and c_0 is the speed of light.

This shows that very close range targets that are equivalent to the pulse width will be distorted and possibly undetected. The typical value of 1 μs pulse width for short-range radar corresponds to a minimum range of about 150 m. The longer pulse widths increase the minimum detectable range. Here are some examples of minimum detectable ranges for different pulse widths and applications:

Typical pulse width τ for:
Air-defense radar: up to 800 μs ($R_{\min} = 120$ km)
ATC air surveillance radar: 1.5 μs ($R_{\min} = 250$ m)
Surface movement radar: 100 ns ($R_{\min} = 25$ m)

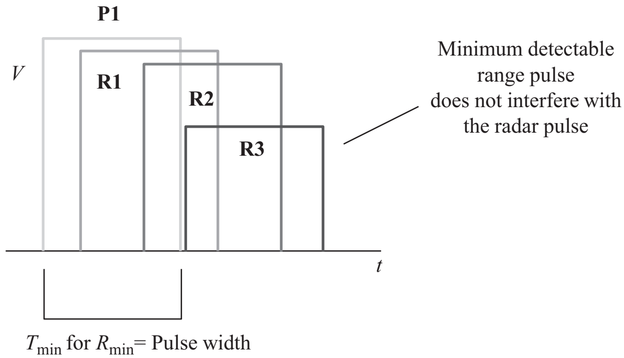


Figure 14.18 Minimum detectable range

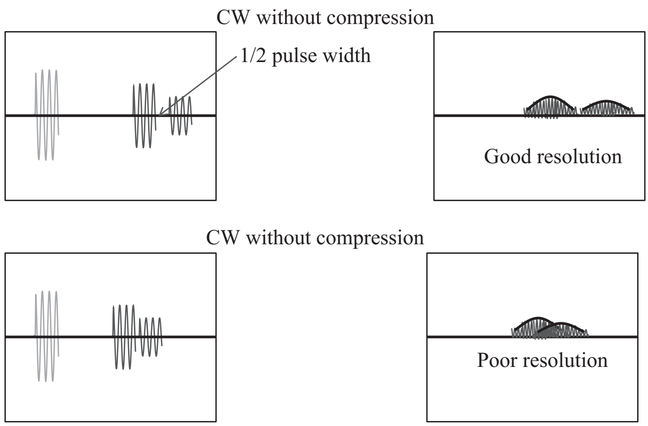


Figure 14.19 Basic radar range resolution

14.3.4.6 Range resolution

Range resolution is the ability to separate two equal targets at the same bearing but different ranges. This depends on many factors such as width of the transmitted pulse, types and sizes of targets, efficiency of the receiver, indicator display, and applications. Weapons-control radar requires high resolution within meters apart. Search radar requires lower resolution within hundreds of meters or miles apart.

Pulse width is the primary factor in range resolution. Generally, the radar is able to distinguish targets separated by one-half the pulse width (Figure 14.19). The theoretical range resolution is equal to:

$$S_r = (c_0 \times \tau)/2$$

where S_r is the range resolution as a distance between the two targets, c_0 is the speed of light, and τ is the transmitters pulse width.

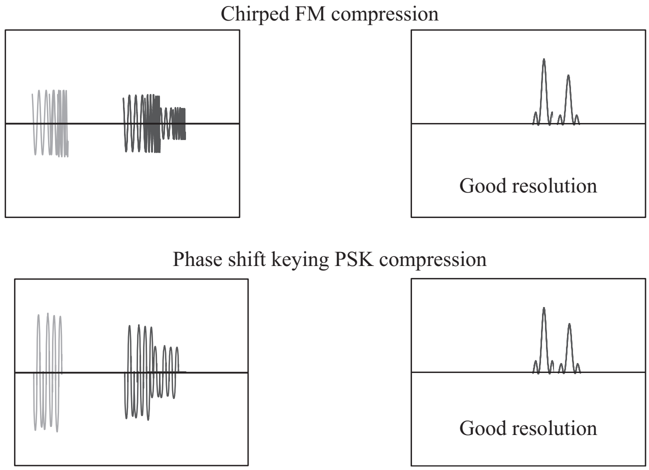


Figure 14.20 Pulse compression improves range resolution using spreading techniques

A valuable technique to increase the range resolution is called pulse compression range resolution. In pulse compression, the minimum range resolution is given by the bandwidth of the transmitted pulse (B_{tx}), not by its pulse width

$$S_r = c_0/2B_{tx}$$

where S_r is the minimum range resolution as a distance between the two targets, c_0 is the speed of light, 3×10^8 m/s, and B_{tx} is the bandwidth of the transmitted pulse.

Pulse compression allows very high resolution using long pulses with higher average power (Figure 14.20). For example:

Given: $B_{tx} = 20$ MHz
Minimum range resolution $S_r = 3 \times 10^8$ m/s/(2 × 20 × MHz) = 7.5 m

Basically, pulse compression improves range resolution by using spreading techniques (Figure 14.20). Both the chirped FM and PSK provide pulse compression using spreading and dispreading of the pulse waveform to narrow the pulses in order to separate them during detection. This provides improved range resolution between different returns.

14.3.5 Bearing

The Angle of Arrival (AoA) is the direction to target and is called the bearing. The bearing is determined by using a directional antenna and is calculated for both azimuth and elevation angles of the directional antenna. The angular measurement accuracy is determined by the antenna's directivity or the gain of the antenna. The greater the directivity or gain of the antenna, the narrower the beamwidth. The narrower beamwidth produces a higher accuracy measurement.

The true bearing is the angle between true north and a line pointed directly at the target (Figure 14.21). The true north reference is measured in the horizontal plane and the true bearing is measured from the true north reference in a clockwise and upward direction from true north for a positive bearing in azimuth and elevation.

The relative bearing uses the direction of the platform such as a ship or aircraft as the reference, and the angle of the target is measured from that reference to the line pointed directly at the target (Figure 14.21). This is also measured in a clockwise and upward direction from the platform reference for a positive bearing in azimuth and elevation.

Bearing is used for tracking targets and is related to the boresite or the position of the target. In nautical navigation, the relative bearing of an object is the clockwise angle in degrees from the heading of the vessel from the observation station on the vessel to the target or from the centerline of the platform as discussed before.

The bearing information is used by servo systems which count the Azimuth-Change Pulses or Elevation-Change Pulses used in older radar antennas and missile launchers. The newer radars use electronic beam pointing such as Active Electronically Scanned (Steerable) Array (AESAs) which replaces older servo systems.

14.3.5.1 Angle resolution

Angle resolution is the ability to separate two equal targets at the same range but different bearings. This depends on many factors such as beamwidth, range, types

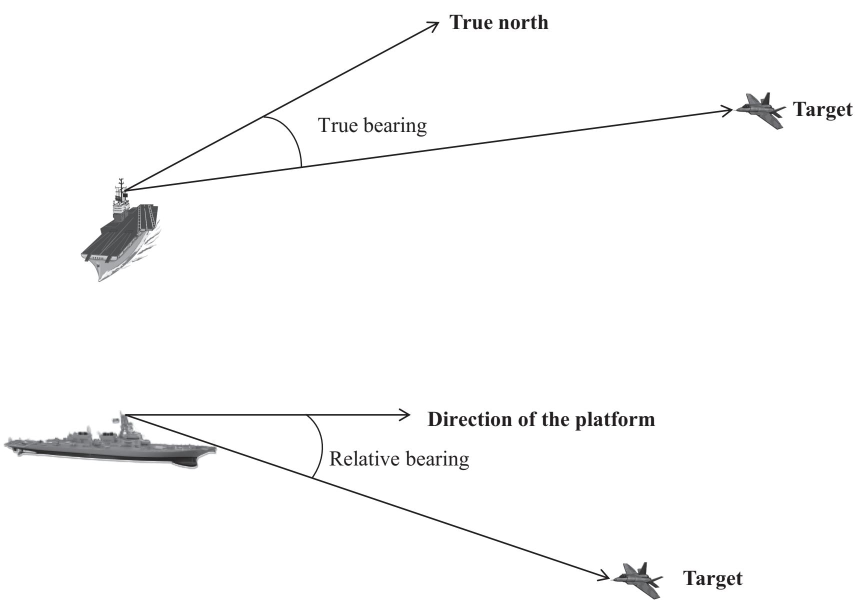


Figure 14.21 True bearing vs relative bearing

and sizes of targets, efficiency of the receiver, indicator display, and applications. Angle resolution is determined by using the 3 dB antenna beamwidth or 1/2 power (Figure 14.22). The targets are resolved if separated by more than antenna beamwidth. Therefore, a narrower antenna beam provides more directivity resulting in better angle resolution. Angle resolution depends on slant range and the equation is shown below for a directional antenna beam for the minimum angle resolution:

$$S_A = 2 \times R \times \sin(\theta/2) \text{ m}$$

where S_A is the minimum angular resolution as a distance between the two targets, R is the slant range, and θ is the 3 dB antenna beamwidth.

For example:

Given: $R = 1{,}000$ m, $\theta = 3°$
Angular resolution $S_A = 2 \times 1{,}000 \times \sin(3/2) = 52$ m

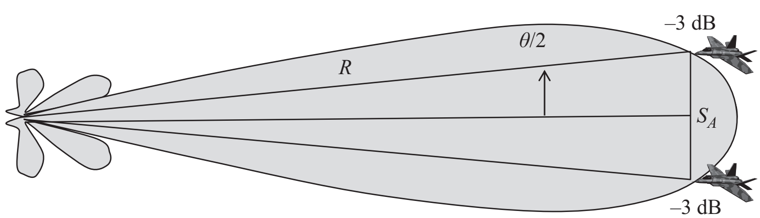
14.3.5.2 Resolution cell

A resolution cell is created using the angle (both azimuth and elevation) and range resolution results (Figure 14.23). This is the basic resolution cell and is formed by the range and angular resolutions. The radar cannot distinguish any two targets that are located inside the same resolution cell. The cell size can be reduced by either shortening the pulse which increases spectrum or by narrowing the aperture angle which reduces interference between the two targets. Therefore, targets located inside the resolution cell cannot be distinguished from each other.

The 3-dB beamwidth of a parabolic antenna is approximated by the equation below:

$$\theta = \sin^{-1}(\lambda/D) = 3 \text{ dB beamwidth}$$

where λ is the wavelength and D is the diameter of the parabolic antenna.



$S_A \geq 2 \times R \times \sin(\theta/2)$ meters

S_A = angular resolution as a distance between the two targets
R = slant range
θ = 3-dB antenna beamwidth

Figure 14.22 Angle resolution for directional beam antenna

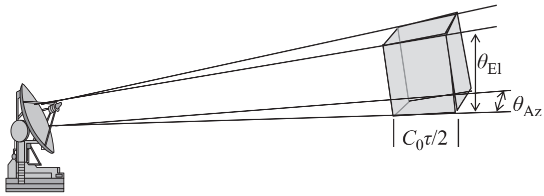


Figure 14.23 Resolution cell determines detection between targets

For example, calculating the 3 dB beamwidth:

Given: $f = 2.4$ GHz, $D = 3$ m
$\theta = \sin^{-1}((c_o/f)/D) = \sin^{-1}((3 \times 10^8)/2.4 \text{ GHz})/3 = 2.39°$

Given $R = 1{,}000$ m
Angular resolution $= 2 \times 1{,}000 \times \sin(2.39/2) = 41.7$ m

14.3.6 Radar accuracy

Accuracy is the difference between the estimated position and the measured position. Radio navigation performance accuracy is a statistical measurement using a Gaussian distribution. The probability that the measurement is within 2 sigma of the estimated position is equal to 95%. This assumes that all known corrections are taken into account so the mean = 0. Any residual bias is small compared with the stated accuracy requirement. The true value is the observed value over a time interval taking into account siting and exposure.

14.3.7 Plan position indicator and A-Scope

There are many ways to display the radar returns for positioning and many types of displays. Two of the most common displays for radar target returns are the Plan Position Indicator (PPI) and the A-Scope. For the PPI, the return is displayed on a circle indicating the angle of the target from 0° to 360° (Figure 14.24). The rotating search radar antennas illuminate the targets on the PPI according to the angle received.

The range is displayed as a line on the PPI display with the target return appearing as a bright blip on the range line at the measured angle. The range of the target is displayed according to the distance from the center of the PPI. The range gate is moved to the location of the target return and the range gate is locked to that point on the PPI and tracks the movement of the target.

The A-scope is used to see the signal return of the target and shows a range gate that is moved to the location of the target return and locks onto that return (Figure 14.24).

14.3.8 Probability of detection and false alarms

Pulsed systems, such as pulse position modulation (PPM) systems and radars, use probability of detection and probability of false alarms since data are not sent on a

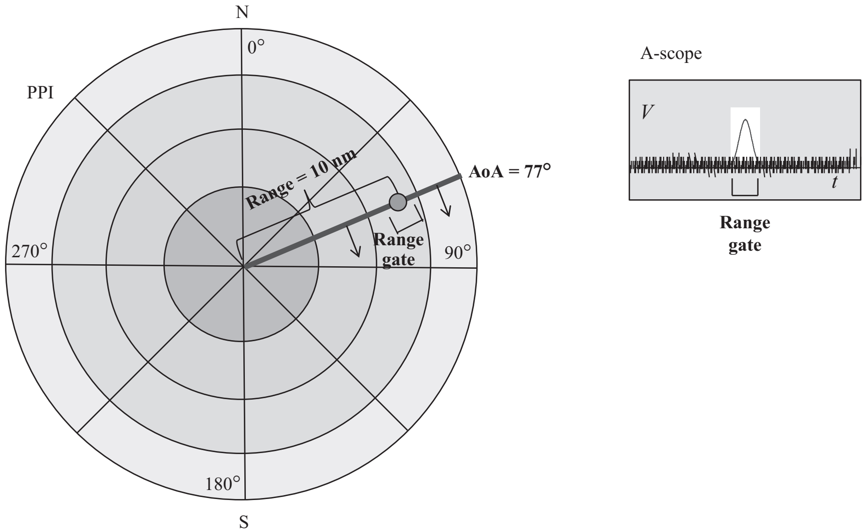


Figure 14.24 Position plan indicator and A-scope displays the radar returns

continuous basis. They operate on whether or not a pulse was detected correctly and whether or not noise or other signals caused the system to detect a signal that was not the true signal (false alarm). It is not just a matter of whether a zero or a one was sent but whether anything was sent at all. Therefore, errors can come from missed detection or false detection.

The time between false alarms (T_{fa}) is dependent on the integration time, which determines the number of pulses integrated (N_{pi}). The number of pulses integrated is divided by the probability of false alarm (P_{fa}) as shown:

$$T_{fa} = N_{pi}/P_{fa}$$

Thus, the longer the integration time, the fewer false alarms will be received in a given period of time.

For a pulsed system, there will be times when nothing is sent, yet the detection process needs to detect whether or not something was sent. There is a threshold set in the system to detect the signal. If nothing is sent and the detection threshold is too low, then the probability of false alarms or false detections will be high when there is no signal present. If the detection threshold is set too high to avoid false alarms, then the probability of detecting the signal when it is present will be too low. If the probability of detection is increased to ensure detection of the signal, then there is a greater chance that we will detect noise, which increases the probability of a false alarm. A trade-off specifies where to set the threshold. Often times, this threshold is selected at the intersection of the probabilities, however, the threshold needs to be evaluated to select the best threshold for the application. Notice that the probability of detection describes whether or not the signal was

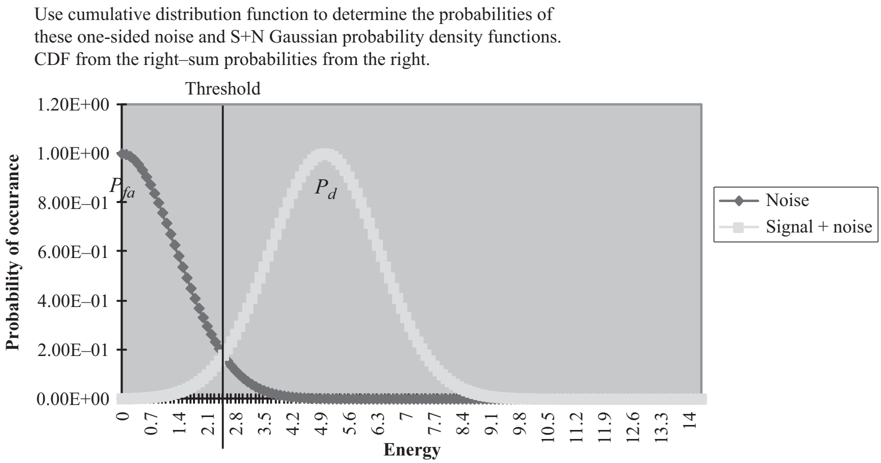


Figure 14.25 Probability of detection and false alarms curves

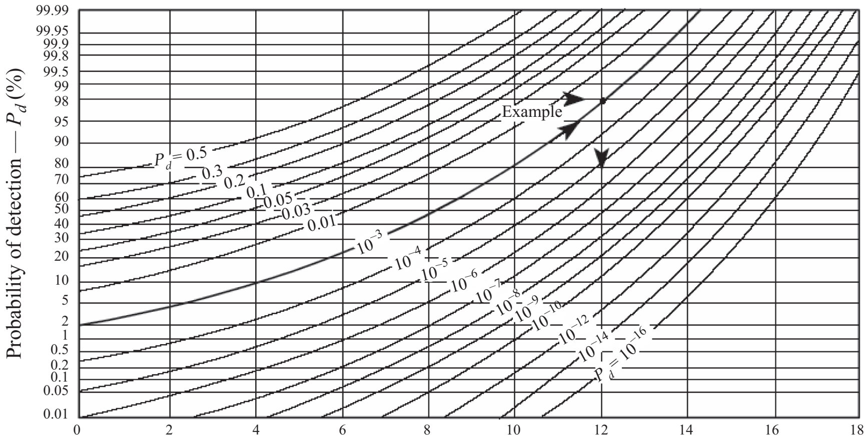
detected. The probability of false alarm refers to whether or not noise was detected. This should not be confused with whether a zero or a one was detected, which is evaluated using the probability of error. A curve showing the PDFs of one-sided noise and signal plus noise is shown in Figure 14.25.

The energy (E_p) is selected to provide the desired probability of detection and probability of false alarm. It is contained in the pulse (E_p). The cumulative distribution functions are used to calculate these parameters. For example, if the threshold is selected at a value of 2.4 (Figure 14.25), then the cumulative distribution function from 2.4 to $+\infty$ (area under the signal) is the probability of detection. The probability of false alarm is the cumulative distribution function from 2.4 to 0 (area under the noise curve from 2.4). This point, where the two Gaussian curves intersect, provides a good point to select, and the trade-offs between detection and false alarms are performed to optimize a design to meet the requirements for the system.

Another useful graph is used to determine what threshold to select (Figure 14.26). The probabilities of detection and false alarms are shown with respect to the level of the SNR. For example, with a 12 dB SNR and a probability of false alarm Pa, the probability of detection Pd is equal to 98%.

The cumulative distribution function for Gaussian distribution can be used since these processes are Gaussian (Figure 14.27). Since the P_{fa} and P_d are integrating the function from 2.4 to $+\infty$, and the total distribution function is equal to unity, then the distribution function is equal to:

$$F_x(x) = 1 - \left(\frac{1}{2}\left[1 + \operatorname{erf}\frac{x}{\sqrt{2}\sigma}\right]\right) = \frac{1}{2}\left(1 - \operatorname{erf}\frac{x}{\sqrt{2}\sigma}\right)$$



Nomograph of S/N as a function of
probability of detection (P_d) and probability of false alarm (P_{fa})

Figure 14.26 Probability of detection and false alarms vs SNR

The complementary error function is equal to:

$$\text{erfc} = 1 - \text{erf}$$

Substituting the erfc in the previous equation produces:

$$F_x(x) = \frac{1}{2}\left(\text{erfc}\frac{x}{\sqrt{2}\sigma}\right)$$

The variance for both P_{fa} and P_d is:

$$\text{Var} = 1/2(E_b/N_o)$$

The standard deviation or σ is the square root of the variance:

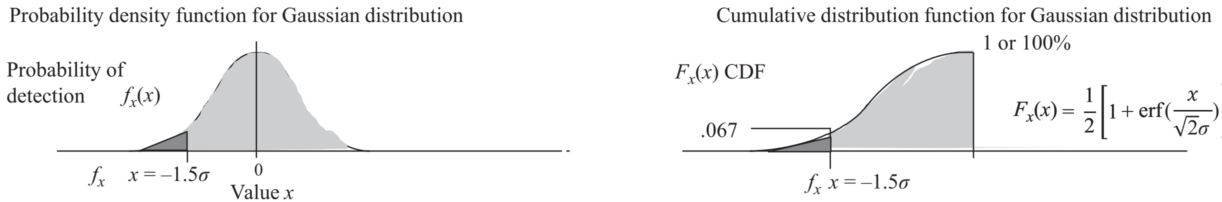
$$\text{StdDev} = \sqrt{\frac{E_b}{N_o}\frac{1}{2}}$$

Substituting the standard deviation into this equation gives:

$$F_x(x) = \frac{1}{2}\left(\text{erfc}\frac{x}{\sqrt{E_p N_o}}\right) = P_{fa}$$

Since P_d contains an offset of the actual standard deviation from zero, this offset is included in the equation as follows:

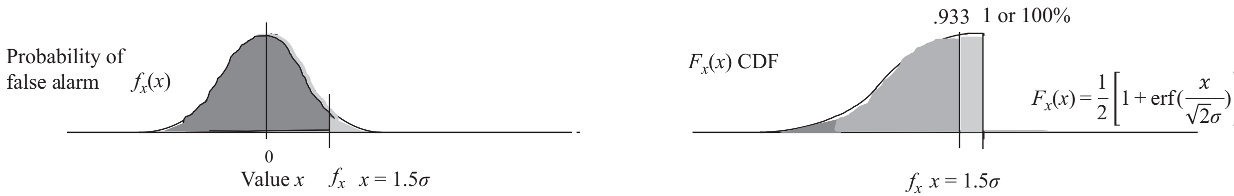
$$F_x(x) = \frac{1}{2}\left(\text{erfc}\frac{x}{\sqrt{E_p N_o}} - \sqrt{\frac{E_p}{N_o}}\right) = P_d$$



$x = -1.5\sigma$

$$F_x(x) = \frac{1}{2}\left[1 + \mathrm{erf}\left(\frac{-1.5}{\sqrt{2}\sigma}\right)\right] = \frac{1}{2}\left[1 + \mathrm{erf}\left(\frac{-1.5}{\sqrt{2}}\right)\right] = \frac{1}{2}\left[1 - \mathrm{erf}\,(1.06)\right] = \frac{1}{2}\left[1 - 0.866\right] = .067$$

Probability inside curve = 1–.067 = .933 = 93.3% (subtract from total)



$x = 1.5\sigma$

$$F_x(x) = \frac{1}{2}\left[1 + \mathrm{erf}\left(\frac{1.5\sigma}{\sqrt{2}\sigma}\right)\right] = \frac{1}{2}\left[1 + \mathrm{erf}\left(\frac{1.5}{\sqrt{2}}\right)\right] = \frac{1}{2}\left[1 + \mathrm{erf}\,(1.06)\right] = \frac{1}{2}\left[1 + .866\right] = .933$$

Probability inside curve = 1–.933 = .067 = 6.7%

Figure 14.27 CDF equation for Gaussian distribution

For P_d, given $x = -1.5\sigma$.

$$F_x(x) = \frac{1}{2}\left[1 + \operatorname{erf}\left(\frac{-1.5\sigma}{\sqrt{2}\sigma}\right)\right] = \frac{1}{2}\left[1 + \operatorname{erf}\left(\frac{-1.5}{\sqrt{2}}\right)\right] = \frac{1}{2}[1 - \operatorname{erf}(1.06)]$$

From Table 14.1: erf(1.06) = 0.866

$$F_x(x) = \frac{1}{2}[1 - 0.866] = 0.067$$

Probability inside curve = 1 − 0.067 = 0.933 = 93.3% (subtract from total)

For P_{fa}, given $x = 1.5\sigma$.

$$F_x(x) = \frac{1}{2}\left[1 + \operatorname{erf}\left(\frac{1.5\sigma}{\sqrt{2}\sigma}\right)\right] = \frac{1}{2}\left[1 + \operatorname{erf}\left(\frac{1.5}{\sqrt{2}}\right)\right] = \frac{1}{2}[1 + \operatorname{erf}(1.06)]$$

$$= \frac{1}{2}[1 + 0.866] = 0.933$$

Probability inside curve = 1 − 0.933 = 0.067 = 6.7%.

P_d is the probability of detecting one pulse at a time. If 10 pulses are required for a system to receive and each of these is independent and has the same probability for detection, then the probability of detecting all 10 pulses is $(P_d)^{10} = 0.5$ or 50%. Error correction is frequently used to provide an increased probability of detection of each of the pulses. In many systems, the probability of missing a pulse needs to be determined and can be calculated by probability theory using the binomial distribution function (BDF).

14.3.9 Pulsed system probabilities using binomial distribution function (BDF)

The BDF is used to calculate the probability of missing a pulse. For example, suppose that the probability of detection is 99% for one pulse. If a message consists of 10 pulses, then the probability of getting all 10 pulses is 0.99^{10} 1/4 0.904, or 90.4%. The probability of missing one pulse is determined by using the binomial theorem as follows:

$$p(9) = \left[{10 \atop 9}\right] p^9 (1-p)^{(10-9)} = \left[{10 \atop 9}\right] (0.99)^9 (0.01)^1$$

$$= \frac{10!}{(10-9)!9!}(0.99)^9(0.01)^1 = 9.1\%$$

Therefore, the percentage of errors that result in only 1 pulse lost out of 10 is

$$\text{Percent (one pulse lost)} = 9.1\%/(100\% - 90{:}4\%) = 95\%{:}$$

If there are errors in the system, 95% of the time they will be caused by one missing pulse. By the same analysis, the percent of the time that the errors in the system are caused by two pulses missing is 4.3%, and so forth, for each of the possibilities of missing pulses. The total possibilities of errors are summed to equal 100%.

Table 14.1 Error function erf(x)

	Hundredths digit of x									
x	**0**	**1**	**2**	**3**	**4**	**5**	**6**	**7**	**8**	**9**
0.0	0.00000	0.01128	0.02256	0.03384	0.04511	0.05637	0.06762	0.07886	0.09008	0.10128
0.1	0.11246	0.12362	0.13476	0.14587	0.15695	0.16800	0.17901	0.18999	0.20094	0.21184
0.2	0.22270	0.23352	0.24430	0.25502	0.26570	0.27633	0.28690	0.29742	0.30788	0.31828
0.3	0.32863	0.33891	0.34913	0.35928	0.36936	0.37938	0.38933	0.39921	0.40901	0.41874
0.4	0.42839	0.43797	0.44747	0.45689	0.46623	0.47548	0.48466	0.49375	0.50275	0.51167
0.5	0.52050	0.52924	0.53790	0.54646	0.55494	0.56332	0.57162	0.57982	0.58792	0.59594
0.6	0.60386	0.61168	0.61941	0.62705	0.63459	0.64203	0.64938	0.65663	0.66378	0.67084
0.7	0.67780	0.68467	0.69143	0.69810	0.70468	0.71116	0.71754	0.72382	0.73001	0.73610
0.8	0.74210	0.74800	0.75381	0.75952	0.76514	0.77067	0.77610	0.78144	0.78669	0.79184
0.9	0.79691	0.80188	0.80677	0.81156	0.81627	0.82089	0.82542	0.82987	0.83423	0.83851
1.0	0.84270	0.84681	0.85084	0.85478	0.85865	0.86244	0.86614	0.86977	0.87333	0.87680
1.1	0.88021	0.88353	0.88679	0.88997	0.89308	0.89612	0.89910	0.90200	0.90484	0.90761
1.2	0.91031	0.91296	0.91553	0.91805	0.92051	0.92290	0.92524	0.92751	0.92973	0.93190
1.3	0.93401	0.93606	0.93807	0.94002	0.94191	0.94376	0.94556	0.94731	0.94902	0.95067
1.4	0.95229	0.95385	0.95538	0.95686	0.95830	0.95970	0.96105	0.96237	0.96365	0.96490
1.5	0.96611	0.96728	0.96841	0.96952	0.97059	0.97162	0.97263	0.97360	0.97455	0.97546
1.6	0.97635	0.97721	0.97804	0.97884	0.97962	0.98038	0.98110	0.98181	0.98249	0.98315
1.7	0.98379	0.98441	0.98500	0.98558	0.98613	0.98667	0.98719	0.98769	0.98817	0.98864
1.8	0.98909	0.98952	0.98994	0.99035	0.99074	0.99111	0.99147	0.99182	0.99216	0.99248
1.9	0.99279	0.99309	0.99338	0.99366	0.99392	0.99418	0.99443	0.99466	0.99489	0.99511
2.0	0.99532	0.99552	0.99572	0.99591	0.99609	0.99626	0.99642	0.99658	0.99673	0.99688
2.1	0.99702	0.99715	0.99728	0.99741	0.99753	0.99764	0.99775	0.99785	0.99795	0.99805
2.2	0.99814	0.99822	0.99831	0.99839	0.99846	0.99854	0.99861	0.99867	0.99874	0.99880
2.3	0.99886	0.99891	0.99897	0.99902	0.99906	0.99911	0.99915	0.99920	0.99924	0.99928
2.4	0.99931	0.99935	0.99938	0.99941	0.99944	0.99947	0.99950	0.99952	0.99955	0.99957
2.5	0.99959	0.99961	0.99963	0.99965	0.99967	0.99969	0.99971	0.99972	0.99974	0.99975
2.6	0.99976	0.99978	0.99979	0.99980	0.99981	0.99982	0.99983	0.99984	0.99985	0.99986
2.7	0.99937	0.99987	0.99988	0.99989	0.99989	0.99990	0.99991	0.99991	0.99992	0.99992
2.8	0.99992	0.99993	0.99993	0.99994	0.99994	0.99994	0.99995	0.99995	0.99995	0.99996
2.9	0.99996	0.99996	0.99996	0.99997	0.99997	0.99997	0.99997	0.99997	0.99997	0.99998
3.0	0.99998	0.99998	0.99998	0.99998	0.99998	0.99998	0.99998	0.99999	0.99999	0.99999
3.1	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999	0.99999
3.2	0.99999	0.99999	0.99999	1.00000	1.00000	1.00000	1.00000	1.00000	1.00000	1.00000

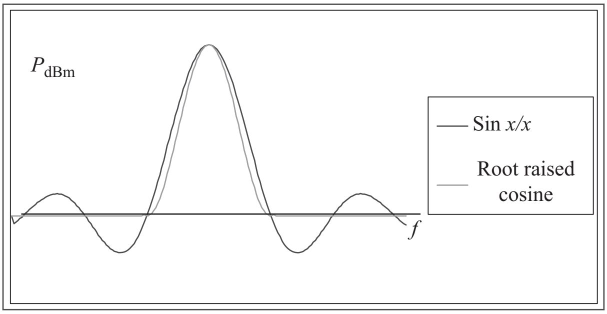


Figure 14.28 Root raised cosine is the recommended pulse shaping

14.3.10 Pulse shaping for radars

Pulse shaping is used on the radar pulse before it is transmitted to the target. The pulse shaping uses different types of filters to produce advantages in the propagation, noise, and detection of the pulse. Pulse shaping can improve the spectrum of a radar system and also can help to reduce sidelobes in a pulse system. A shaped pulse in the time domain creates a more square pulse in the frequency domain so that it does not interfere with other signals especially outside the bandwidth of the radar and improves the detecting of the pulse. A root raised cosine pulse shaping is ideal for providing a pulse with minimum out-of-band power (Figure 14.28). The disadvantage of a shaped pulse is reduced resolution since the pulses are rounded. Therefore, there is a trade-off between out-of-band transmissions and resolution that is dependent on the application of the radar.

14.4 Clutter

Clutter is unwanted RF returns from objects in the radar range and space. It interferes with the desired return and causes false alarms of a desired signal. The clutter is a return of other objects other than the target. It also can be a jammer that monitors the radar signal and sends back a return that appears like a target but causes distortion of the desired target. There are many types of clutter which include: ground, mountains, sea, atmospheric turbulence, ionosphere, birds, buildings, cars, metal objects, and chaff from countermeasure devices. Also clutter can be anomalies in hardware, self-generated noise, rotating antennas, and colocated devices and equipment.

Clutter can be mitigated by using a Moving Target Indicator (MTI) since most clutter is stationary. The MTI method uses the principle of moving targets that generate Doppler and can be distinguished from stationary targets that do not generate Doppler. Doppler processing uses filters to separate clutter from desirable signals. Sea clutter can be reduced by using horizontal polarization of the antennas. In addition, circular polarization can reduce rain clutter.

The standard deviation of clutter power spectrum is equal to:

$$\sigma_c = 2\sigma_v/\lambda$$

where σ_v is the clutter standard deviation in meters/second.

Long runs of waveguide between the radar and antenna can cause a phenomenon called sunburst. Sunburst appears at the center of the PPI display as the RF reflects off dust or misguided signals in the waveguide. This can be reduced by adjusting the timing between transmitter pulse and received echo.

14.5 Radar frequency bands

Radars use many different frequency bands for operation. Radar systems generate the pulse and then transmit electromagnetic radio waves at the frequency selected. The reflected radio waves are received and detected by the radar receiver. The frequency of the radio waves used depends upon the radar application and design. The required antenna size is proportional to wavelength and inversely proportional to frequency and the antenna gain is dependent on antenna size or area of the reflector and the frequency or wavelength. Large antennas have more gain and the beam is more tightly focused at a given frequency which provides better directivity. Most airborne radars operate between L- and Ka-bands and some commonly used radar frequencies are listed in Table 14.2.

Many short-range targeting radars, where the platform is a tank or helicopter, operate in the millimeter band between 40 and 100 GHz. Many long-range ground-based operate at UHF or lower frequencies. Radars select the frequency of operation for their individual requirements and applications including the ability to use large antennas for increase gain and to reduce the freespace loss, atmospheric attenuation, and ambient noise.

For example, lower frequency radars are used for lower freespace loss, larger antennas for higher gain and long-range search applications, and easier to produce higher PAs. In addition, since the wavelength is longer compared to the target extent, the targets are Rayleigh scatterers which have small, nonfluctuating RCS.

Table 14.2 Radar commonly used frequency bands

Radar band	Frequency (GHz)	Wavelength (cm)
Millimeter	40–100	0.75–0.30
Ka	26.5–40	1.1–0.75
K	18–26.5	1.7–1.1
Ku	12.5–18	2.4–1.7
X	8–12.5	3.75–2.4
C	4–8	7.5–3.75
S	2–4	15–7.5
L	1–2	30–15
UHF	0.3–1	100–30

Another advantage of using low frequency radars is the increased range achieved by reflecting off the ionosphere. A list of advantages and disadvantages for several different frequency bands are listed in Table 14.3.

There are many applications and uses for all of these bands and the applications increase over time. A list of the typical applications in use today is in Table 14.4.

Note that the bandwidth is what determines the range resolution and frequency agility capabilities of the radar.

Another factor that is important in the design of the radar hardware is to maintain constant group delay. This prevents distortion and rounding of the pulses in the receiver and maintains the square pulse shape needed to prevent resolution degradation.

14.6 Moving target indicator

The purpose of using a MTI is to reject signals from fixed or slow-moving unwanted targets such as buildings, hills, trees, sea, rain, and debris and to detect and display the desired moving targets such as aircraft, automobiles, and ships. MTI separates moving targets from stationary targets and reduces or eliminates the stationary (nonmoving) returns to the radar. An example of the usefulness of the MTI is for moving target such as an aircraft flying against a mountain which would contain large amounts of nonmoving target echoes. However, with MTI, these are rejected and the moving aircraft can then be detected. The most common design of

Table 14.3 Advantages and disadvantages of different radar frequency bands

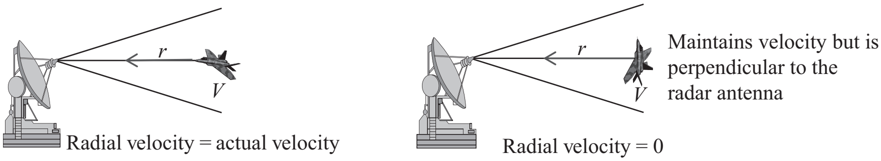
Radar bands	Advantages	Disadvantages
A & B-bands HF&VHF below 300 MHz	Easier to get HPAs, low freespace loss, max range	Large antennas, low angle accuracy & angle resolution
C-band (UHF-radar)	Not affected by clouds and rain, smaller antennas	Increase in freespace loss
D-band (L-band radar)	Broad bandwidth	Limited to LOS range
E/F-band (S-band radar)	Smaller antenna, lower range	Higher atmospheric loss, need more power
G-band (C-band radar)	Smaller antenna size, better accuracy and resolution	Atmospheric turbulence, use circular polarization
I/J-band (X- and Ku-band radars)	Very small antenna size, good resolution, mobile apps	High freespace loss
K-band (K- and Ka-band radars)	High accuracy and range resolution	Higher atmospheric attenuation, shorter range
Higher bands L/M-band (Q-W-bands)	High accuracy, range resolution, LPI, low cosite jamming	Very high loss, very short-range

Table 14.4 Applications and uses for different radar frequency bands

Radar bands	Applications
A- and B-band (HF- and VHF-radar) below 300 MHz	Early warning radars and over the horizon radars Communications and broadcasting services with limited BW but high accuracy and resolution
C-band (UHF-radar)—300 MHz to 1 GHz	Detecting/tracking satellites and ballistic missiles over a long-range Early warning and target acquisition, surveillance, weather Ultrawideband uses A-C bands, Ground Penetrating Radar for archaeological explorations
D-band (L-band radar)—1 to 2 GHz, long-range 250 NM air-surveillance radars	Air Traffic Management long-range surveillance radars Air Route Surveillance Radar Monopulse Secondary Surveillance Radar
E/F-band (S-band radar)	Special Airport Surveillance Radars used at airports, of 50–60 NM
G-band (C-band radar)	Mobile military battlefield surveillance, missile-control, & ground surveillance
I/J-band (X- and Ku-band radars)—8–12 GHz	Military airborne applications for interceptor, fighter, and attack of enemy fighters and of ground targets Missile guidance systems Maritime civil and military navigation radars Space borne or airborne imaging radars—Synthetic Aperture Radar (SAR) for military electronic intelligence and civil geographic mapping Inverse Synthetic Aperture Radar in maritime airborne instrument of pollution control
K-band (K- and Ka-band radars)	Surface movement radar or airport surface detection equipment
Higher bands L/M-band (Q-W-bands)	Radar applications limited for a short-range, a few meters Automotive, parking assistants, blind spot, and brake assists Future 75 and 96 GHz

an MTI is to use the Doppler Effect created by moving targets. In addition, the MTI radar uses a low PRF to avoid range ambiguities. Often time two different MTI terms are used depending on their applications; Airborne MTI—AMTI and Ground MTI—GMTI.

MTI is based on the movement of the desired target with a minimum movement required to separate it from the stationary targets. This is known as Minimum Detectable Velocity (MDV) (Figure 14.29). This is the radial component of a target's velocity, and as it approaches zero, the target will fall into the clutter or blind zone. This can occur by either slowing the aircraft down as it approaches the radar antenna until it approaches zero, or for a more practical scenario the aircraft becomes perpendicular to the radar antenna which produces zero radial velocity (Figure 14.29). Any target with a velocity less than MDV cannot be detected since



$$\text{MDV} = \frac{\lambda}{2}\left(\frac{4v_p}{B}\sqrt{(\sin \text{Az} \times \sin \text{El})^2\,(\cos \text{Az} \times \cos \text{El})^2}\right)$$

Figure 14.29 Minimum detectable velocity (MDV)

there is not sufficient Doppler shift in its echo to separate it from the main lobe clutter return. The MDV is calculated as follows:

$$\text{MDV} = \lambda/2\left\{4V_p/B\left[(\sin \text{Az} \times \sin \text{El})^2 + (\cos \text{Az} \times \cos \text{El})^2\right]^{1/2}\right\}$$

14.7 One-way passive radar Doppler effects

Passive radar transmits in only one direction and there is no reflection or return. The Doppler effects the carrier frequency of one-way transmission for the passive radar. The target sends a radar signal to the radar receiver. As the target moves in the direction of the radar (radial velocity), the distance to radar changes and the frequency changes depending on the direction of the target. The frequency increases as the target moves closer and the frequency decreases as the target moves farther away. The change in frequency equals the Doppler frequency and is calculated as follows:

$$f_{\text{Doppler}} = (v_r/c_o) \times f_0 = v_r/\lambda_0$$

where f_{Doppler} is the Doppler frequency, v_r is the radial velocity, f_0 is the radar frequency, c_o is the speed of light, and λ_0 is the wavelength of radar frequency.

For example:

Given: $v_r = 100$ km/h $= 27.78$ m/s
$f_0 = 2.4$ GHz
$f_{\text{Doppler}} = (27.8 \text{ m/s}/3 \times 10^8 \text{ m/s}) \times 2.4 \text{ GHz} = 222$ Hz

14.8 Two-way active radar Doppler effects

Active radar is the basic radar that sends out a pulse and receives the reflected pulse or echo from the target. This represents a two-way path from the radar to the target, then from the target to the radar. The radar transmits the pulse frequency to the

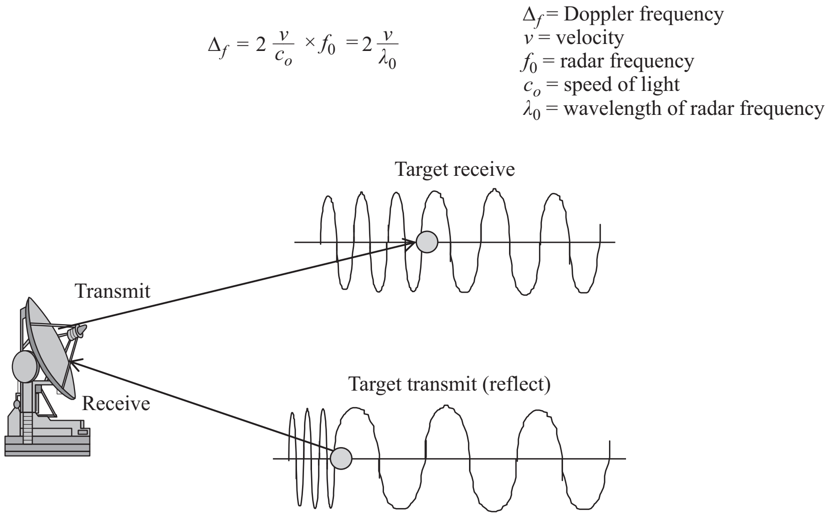


Figure 14.30 Active radar Doppler—two way

moving target which causes a Doppler shift on the path to the target. The target reflects the signal including the Doppler frequency back to the radar but still is moving in the direction of the radar which causes an additional Doppler shift on the return. This doubles the effect of the target moving toward the radar so the Doppler shift is twice the frequency (Figure 14.30). This shows that the frequency increases as the target moves closer to the radar and the frequency decreases as the target moves farther away from the radar. To illustrate the concept better, the radar is on the moving aircraft and the parabolic antenna is the target. The change in frequency equals the Doppler frequency and the equation for the active radar is:

$$f_{\text{Doppler}} = 2(v_r/c_o) \times f_0 = 2v_r/\lambda_0$$

where f_{Doppler} is the Doppler frequency, v_r is the radial velocity, f_0 is the radar frequency, c_o is the speed of light, and λ_0 is the wavelength of radar frequency.

For example:

Given: $v_r = 100$ km/h $= 27.8$ m/s
$f_0 = 2.4$ GHz
$f_{\text{Doppler}} = 2(27.8 \text{ m/s}/3 \times 10^8 \text{ m/s}) \times 2.4 \text{ GHz} = 444$ Hz

14.9 MTI sampling process

A simple sampling process and detection is used to eliminate the stationary target and detect the moving target. A radar pulse is sent from the radar to a stationary target and it is reflected back with two echoes (Figure 14.31). A second radar pulse is transmitted and two echoes are again reflected back, however, the first returns

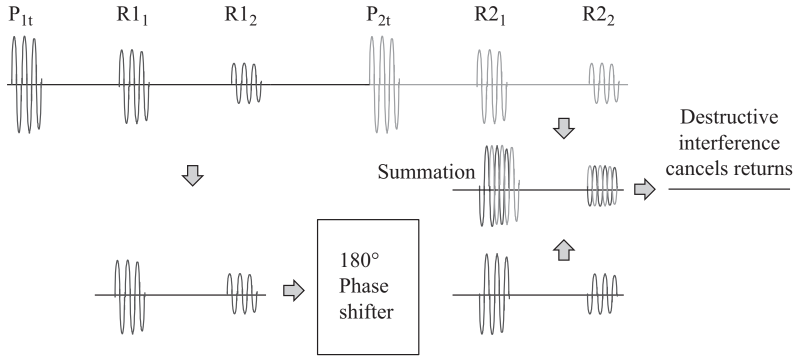


Figure 14.31 Destructive interference cancels stationary targets

went through a 180° phase shifter. Then the two sets of returns are lined up in time and are summed together. Since the stationary targets did not move, the phase and frequency of both sets of returns are the same, so with an 180° phase shift of the first returns, they will cancel each other out (Figure 14.31). This is known as destructive interference.

The returns of the same radar example are used in a moving target scenario. This shows the same setup only this time the phase/frequency do not cancel out due to the change in frequency/phase due to the moving target. In this case, the frequencies experience constructive interference and the returns are amplified (Figure 14.32). This shows the best case where the Doppler has exactly 180° phase shift due to Doppler shift. In a typical configuration, the signal is larger than the stationary target that was canceled and therefore the moving target was detected.

The I/Q detector for Doppler-shifted pulses is shown in Figure 14.33. Both I and Q channels show the detected pulse amplitude depending on the phase for each pulse. Connecting all of the pulses in a continuous manner traces out the Doppler frequency. Note that when the I channel Doppler frequency is at its highest point, the Q-channel Doppler frequency is at its lower point.

An A-scope shows the different displays for the stationary targets and the moving targets (Figure 14.34). The stationary target is stable and does not change. The moving target is moving up and down due to the movement of the target. This creates an amplitude modulated target and can be separated from the stationary target.

Phase jitter, Doppler effects, and environmental influences limit the effectiveness of the MTI process. High power microwave devices are not phase-stable and phase errors between 1st and 2nd pulse cause misalignment of the pulses. With these types of phase errors, the MTI process may not cancel stationary targets. Often times the 180° phase shifter needs to be adjusted to compensate for phase jitter.

Another phenomenon that occurs in MTI radars is at certain speeds the moving target are actually eliminated or reduced. These speeds are called

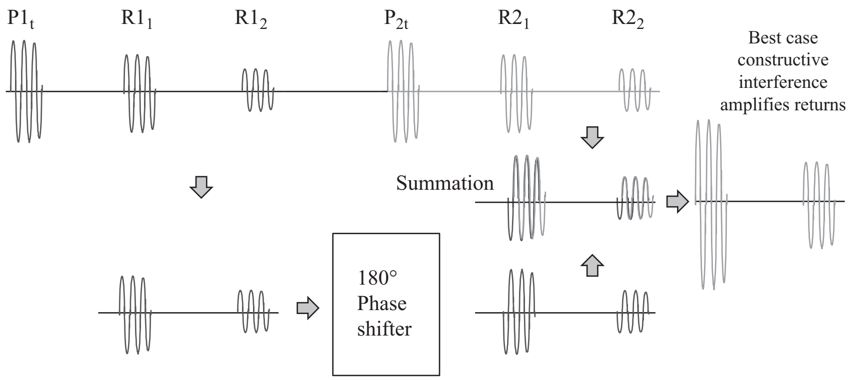


Figure 14.32 Constructive interference amplify moving targets

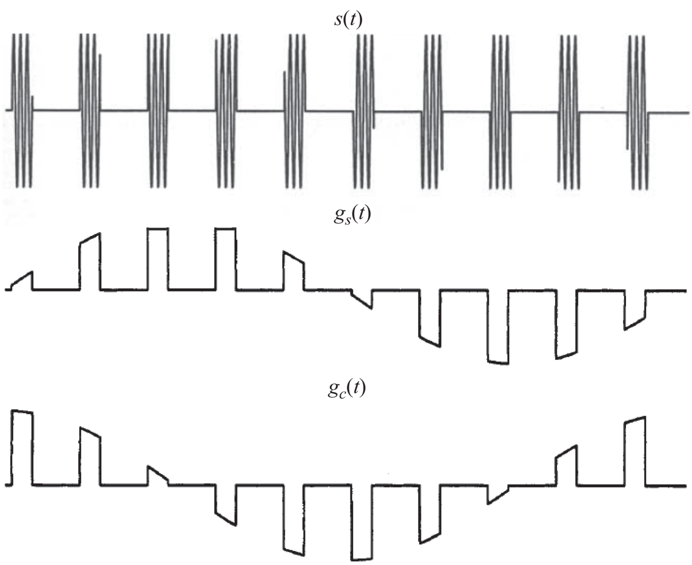


Figure 14.33 Doppler-shifted pulses using an I/Q detector

blind speeds (Figure 14.35). The equation to determine the blind speeds is as follows:

$$v_b = k\lambda f_{\text{PRF}}/2$$

where v_b is the blind speed where the moving target is canceled, k is the an integer value, 1,2,..., λ is the wavelength of the carrier, and f_{PRF} is the PRF of the radar pulses.

For example:

Given: $k = 1, 2, 3, f_o = 2.4, \lambda = 0.125, f_{\text{PRF}} = 500$ Hz

$v_b = 1 \times 0.125 \text{ m} \times 500 \text{ Hz}/2 = 31.25 \text{ m/s}$

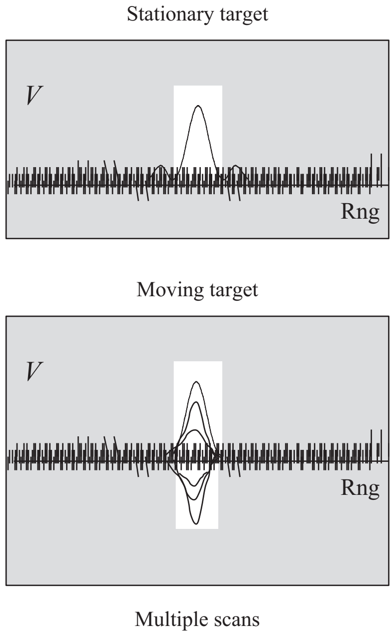


Figure 14.34 A-scope display of stationary and moving targets

$v_b = 2 \times 0.125 \times 500\ \text{Hz}/2 = 62.5\ \text{m/s}$
$v_b = 3 \times 0.125 \times 500\ \text{Hz}/2 = 93.75\ \text{m/s}$

Here is an example of the blind speed and Doppler calculations.

Given: $v_b = 75$ m/s, PRI = 2 ms, PRF = 500 Hz, $c_o = 3 \times 10^8$, $f_o = 1$ GHz

One-way Doppler $= f_{\text{Doppler}} = (v_b/c_o) \times f_o = (75/3 \times 10^8)1 \times 10^9 = 250$ Hz
Two-way Doppler $= f_{\text{Doppler}} = 2(v_b/c_o) \times f_o = 2(75/3 \times 10^8)1 \times 10^9 = 500$ Hz
Doppler = (500 cycles/s × 360°/1 cycle) = 180,000°/s
MTI produces 180,000°/s × 2 ms = 360° = 0° = appears as no phase shift

The response is the same as stationary targets so the moving target is not detected.

14.10 Multiple pulse MTI radar

Multiple radars at different frequencies can help to mitigate the blind speeds. Using two different frequencies causes the blind speeds to be at different locations and when combined together eliminates the blind speed cancelation problem in MTI, see Figure 14.36.

Another method that can be used to mitigate the blind speed problem in MTI is to incorporate multipulse strategies using staggered pulses with irregular PRIs. This changes the PRF so that the blind speed is constantly changing and prevents staying

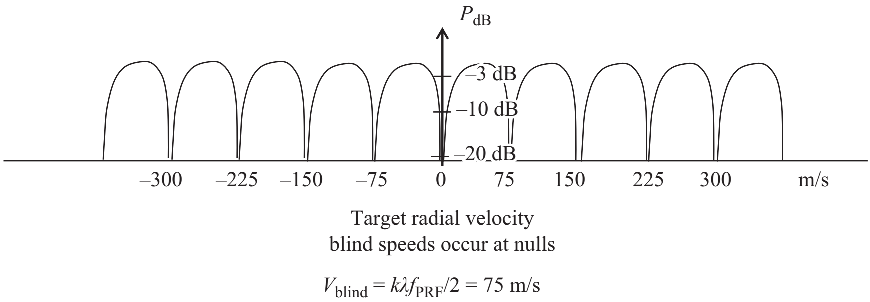


Figure 14.35 Blind speeds for moving targets appear as stationary targets

in a blind speed null. MTI requires 3 or 4 pulses to reduce the effect of blind velocities. MTI detects moving objects 300 times smaller in close proximity to larger stationary objects. Pulse-Doppler signal processing is required to achieve greater subclutter visibility.

14.11 Types of radar antennas

There are basically three types of antennas that are used in radar. They are omnidirectional, reflector mirror antennas, and AESAs. There are advantages and disadvantages to these types of radar antennas as shown in Table 14.5.

The omnidirectional antenna is a vertical monopole antenna with a length of $\lambda/4$ and uses ground plane for an effective $\lambda/2$ length antenna (Figure 14.37). Loaded coils can be used to reduce actual physical length of the monopole antenna while still maintaining the electrical length needed for the optimal performance of the omnidirectional antenna.

There are many types of reflector mirror antennas (Figure 14.38). The most common type is a parabolic reflector which is used for all types of applications including radar. The parabolic shape of the antenna focuses the beam to one point. The larger the surface of the parabolic reflector or dish, the more focused the energy which produces more antenna gain. For the parabolic reflector antenna, the signal is reflected so that all of the fields travel in near parallel paths toward the target. The parabolic surface ensures all paths from surface to feed are equal and are focused toward the target. The parabolic reflector antenna is most often used for high directivity applications.

The other reflector mirror antennas are subsets of the parabolic antenna. The Truncated Paraboloid is a parabolic antenna that has been cut or shortened vertically (Figure 14.38). Therefore, this antenna provides a fan-shaped wide vertical beam with less gain than the parabolic reflector. This allows use for discovering targets by providing more area coverage in the beam in the vertical direction. This helps to detect aircraft at different altitudes without changing the tilt angle of the

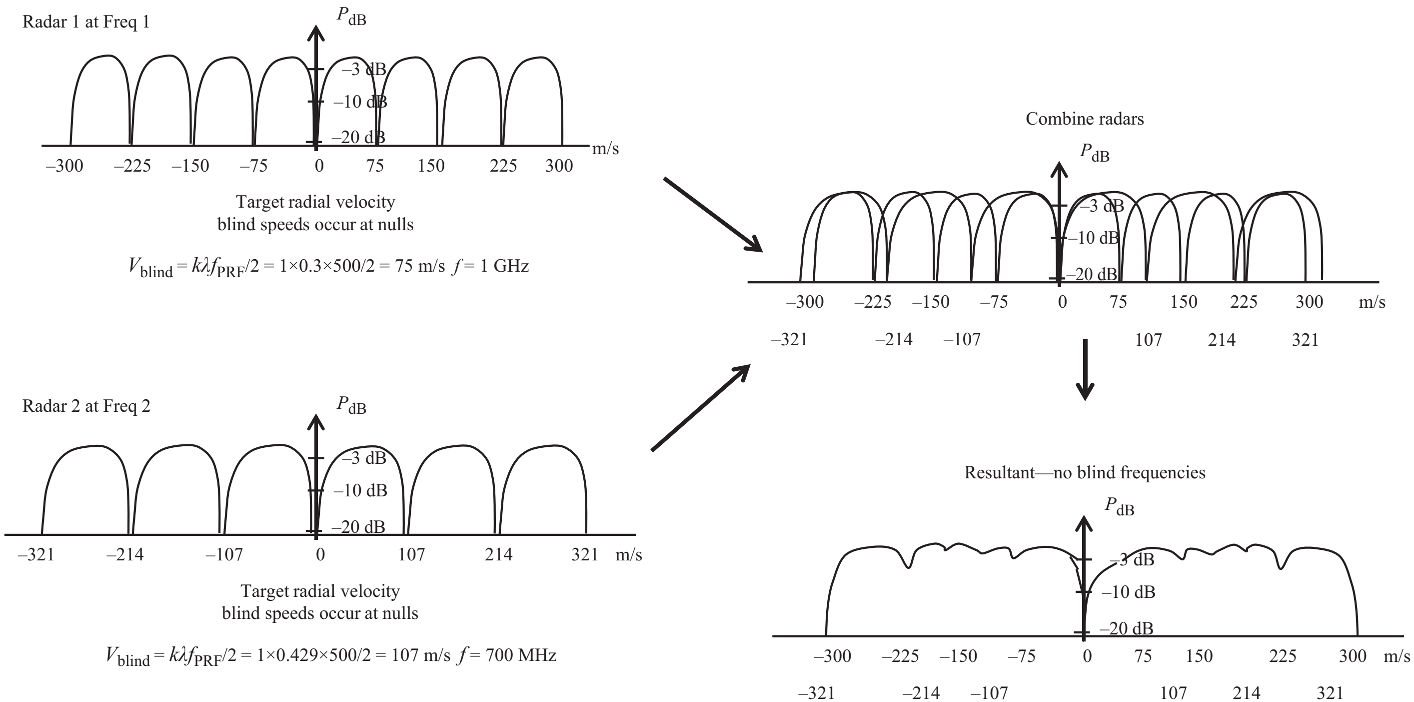


Figure 14.36 Blind speed mitigation using multiple radars at different frequencies

Table 14.5 Advantages and disadvantages of different types of radar antennas

Type of antenna	Advantages	Disadvantages
Omnidirectional	Low cost, small size, 360° instantaneous coverage, low gain	Range only, no bearing information
Reflector mirror	Narrow beam, range and bearing information, 360° slow scanning gimbal coverage, high gain	Medium cost, large size, slow scanning
AESAs	360° very fast scanning coverage (need 4 flat panel arrays), medium size and weight, high gain, multiple beams, versatile	Very high cost, multiple panels needed for 360° coverage, high temperature, complex

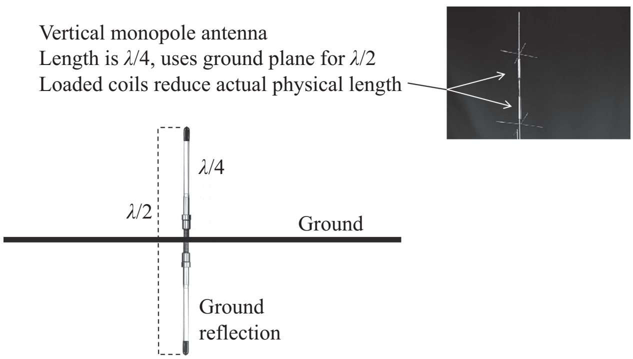


Figure 14.37 Omnidirectional antennas for 360° coverage

antenna. This also can be truncated in the horizontal plane and provides more coverage in the horizontal direction. Therefore, this type of antenna is a parabolic antenna that is cut or truncated in one of the planes, either vertical or horizontal. The Orange Peel Reflector has an orange peel shape like a truncated antenna in the vertical plane and provides a fan-shaped wide vertical beamwidth for searching targets at different elevations. The common use for this type of antenna is during radar acquisition with the antenna rotating in the horizontal plane. The Parabolic Cylinder creates a pattern that is similar to the orange peel reflector by providing narrow vertical line coverage only in the vertical plane (Figure 14.38). It is used to prevent radiation out the ends and is sometimes called a pillbox antenna. The Offset-feed Dish uses a signal feed that is offset to the side of the dish. This helps to prevent blockage of the feed directly in front of the dish. This type of antenna is used in small parabolic dish antenna applications. The Corner-Reflector uses two flat conducting sheets that meet at an angle to form a corner.

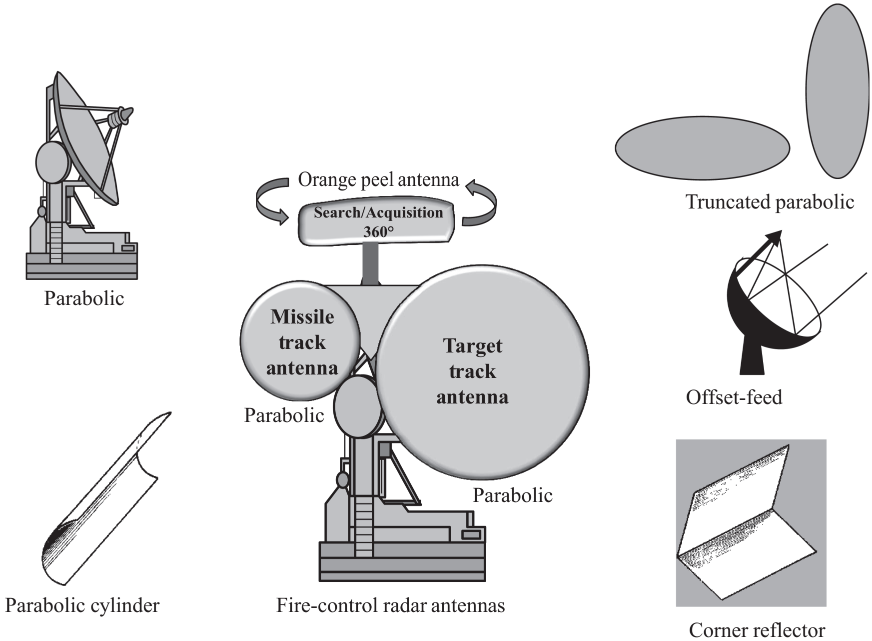


Figure 14.38 Types of directional antennas

The AESA consists of multiple radiating electronic elements. Each of these elements is adjusted in phase and amplitude to produce the antenna pattern. Beam steering can be accomplished without mechanical movement as well as beam shaping including beam spoiling by adjusting the elements in the array. In addition, null steering can be used to move a null in the antenna pattern in the direction of interfering signal to reduce that signal. AESAs also can nearly instantaneously change the pointing direction or reconfigure the beam pattern since it is electronically steered instead of mechanically steered. The AESA is generally a flat panel array and can form beams and multiple beams very quickly (Figure 14.39).

14.12 Block diagrams of a pulse radar system

The radar transmitter block diagram is shown in Figure 14.40. The radar pulse is generated and formatted in the pulse generator module. This includes the pulse width, PRF, pulse shaping, and amplitude. The upconversion module selects the carrier frequency that the pulse will modulate. It uses a LO to select the frequency, and then uses a mixer or other type of modulator to create the 100% AM or on/off keying of the LO. This is then filtered to eliminate any unwanted harmonics or spurious signals that are generated in the modulation process. The RF PA raises the

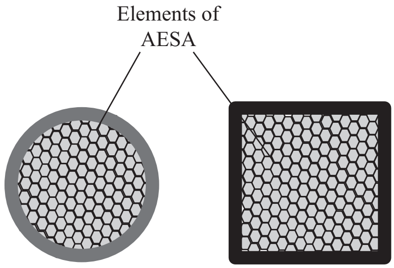


Figure 14.39 Flat panel AESAs

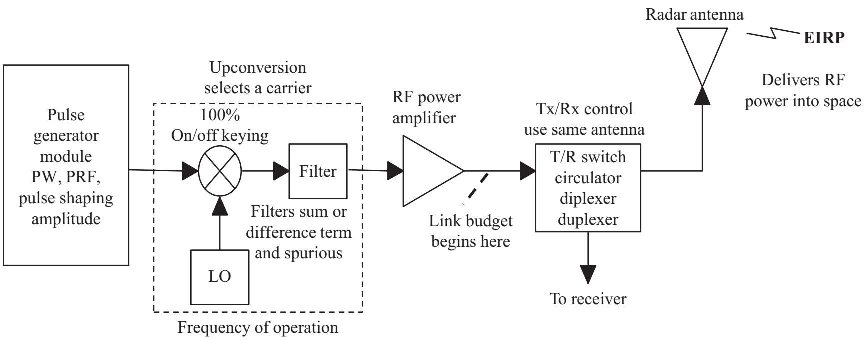


Figure 14.40 Radar transmitter block diagram

power of the modulated signal according to the range of the radar specified in the requirements. The link budget analysis begins at the output of the PA. A Tx/Rx control module allows for use of a single antenna for both transmit and receive. This can be a T/R switch, a circulator, a diplexer, or a duplexer. The output is fed into the radar antenna which provides the antenna gain or directivity for the EIRP into space (Figure 14.40).

The radar receiver block diagram is shown in Figure 14.41. The RF radar waveform is received by the receiver antenna which provides gain by increasing the SNR in the receive path. The losses from the antenna to the first amplifier are included in the link budget and affect the SNR received. The first amplifier is called the low noise amplifier and is used to establish the noise figure for the receiver. Therefore, it is critical to design this first amplifier for minimum noise added to the signal. Implementation losses due to imperfections in the hardware need to be added to the link analysis and the signal then continues through the receiver including the downconverter to remove the carrier frequency and to the detector to detect the timing of the reflected signal or echo from the target. Digital demodulation of a radar pulse is generally used for detection with the minimum distortion of the radar pulse to maximize detection and to minimize false alarms.

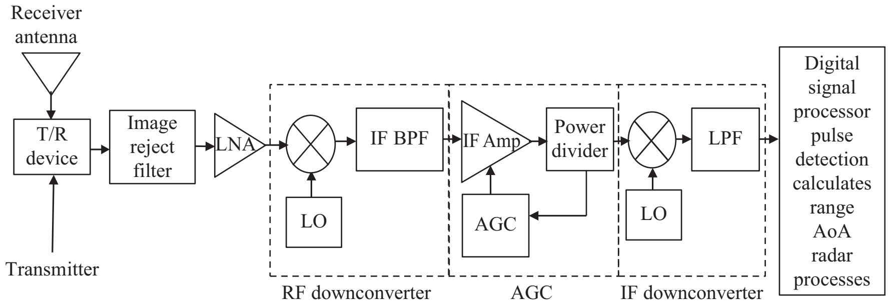


Figure 14.41 Radar receiver block diagram

14.13 Other types of radars

Monostatic radars are the most common radars. These types of radars transmit and receive the signals using the same antenna. Bistatic radars use transmit and receive antennas that are at difference locations separated at a distance approximately equal to expected target range. These types of radars are used for long-range air-to-air and surface-to-air missile systems. Different types of radars will be evaluated in the following pages.

14.13.1 Search, acquisition, and track radar

Search and acquisition and track radars use a search antenna with a wide beamwidth such as an orange peel antenna. This antenna generally uses a mechanical rotating gimbal that the antenna is mounted on and uses a wide elevation beamwidth and scans 360° over the horizon. AESAs can also be used that do not require a gimbal and scans the area electronically. Tracking radars provide continuous positional data on a target from the returns. Most military tracking radar systems are also called fire-control radars. The volume search generally uses other means to direct the radar to the general location of the target. Acquisition searches a small volume of space in a prearranged pattern until the target is found. Once the target is found, the tracking radar locks on to the target using one of several possible scanning techniques. These three modes of operation are used in most fire-control radars and they generally use a high PRF, a narrow pulse width, and a narrow beamwidth for high accuracy and also make it harder for initial target detection. Tracking radars generally focus on individual targets, however, some radars track multiple targets simultaneously. There are two basic types of tracking radars; Conscan and Monopule. Conscan radar has been used for several years and is a very common type radar tracker. Conical scan creates a scan circle around boresite. At boresite, the returned power will be equal at all angles. If the target is off boresite, the power is greater toward the target and less around the rest of the scan circle, with the minimum power in the opposite direction of the target. This creates amplitude modulation of the received power which is used to point the antenna directly at the target. Most radars that have been designed since the 1960s are

Monopulse radar systems. A Monopulse system compares received signal of a single radar pulse against itself. It compares the signal as seen in multiple directions, polarizations, or other differences. The antenna is split up in quadrants with two side-by-side quadrants for Azimuth and two top and bottom for Elevation. It simply takes the sum and difference measurements in the quadrants and calculates the position, see Chapter 10.

14.13.2 Missile and missile guidance radars

The Missile Guidance Radars control and guide a missile to a hostile target. The acquisition and track starts from launch of the missile until detonation. The missiles use the missile guidance radar to intercept targets in three basic ways;

1. Beam-rider missiles follow a beam of the platform's radar energy that is kept continuously pointed at the desired target
2. Homing missiles detect and home in on the platform's radar energy reflected off the target to the missile
3. Passive homing missiles home in on energy that is radiated by the target

Missiles can also be equipped with radar to intercept targets for self-guidance. The missile tracking radars are generally part of a fire-control tracking radar. A 3-in-1 fire-control tracking radar which contains target tracking, missile tracking, and search/acquisition antennas is shown in Figure 14.42. This type of radar system consists of a widebeam or scanning acquisition radar, target track radar to track the target, and a missile track radar to track the missile. The basic idea is that the radar

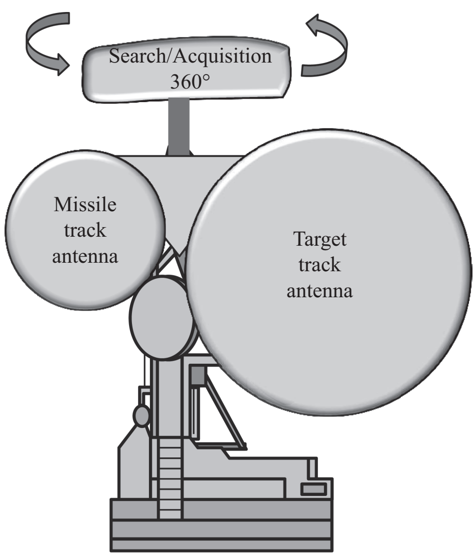


Figure 14.42 3-in-1 fire-control radar includes search/acquisition, target track and missile track

on the ground illuminates the target, the missile homes in on reflected signal off target which can use either pulsed or continuous type radars, air-to-air, air-to-ground, and ground-to-ground applications. Missiles can also contain their own radar signal.

There are many types of missiles including the AM-7 Sparrow, the Standard Missiles (SM-2), the Evolved Sea Sparrow Missiles (ESSM) AARAM, the TOW missile, the HARM missile, and many others.

14.13.3 Airborne radars

Airborne radar is designed for low size, weight, and power. They develop high peak power for the radar pulse with low average power so that the power supplies can be minimal. Antennas are mounted for best coverage and lowest blockage from the aircraft body. The antennas are mounted in domes or pods that form part of the fuselage or attach to the underside of the wings. Often they are mounted in the nose or tail of the aircraft. Fighter aircraft are focused on search, interception, and destruction of enemy aircraft which requires search, acquisition, and tracking radars.

14.13.4 Frequency diversity radar

Frequency Diversity Radar uses two pulses at two different frequencies with multiple radars or antennas. The pulses are radiated one after another at very short intervals. If the frequencies are separated, the echo signals of a fluctuating target are statistically decorrelated. The advantage of this type of radar is that it sums both signals together which achieves a high max signal with 3 dB of gain. This provides a higher maximum range with equal probability of detection and false alarm. It also smoothes the fluctuation of the complex echo signal which is reduce by 3 dB. The reflected single signals must be independent to increase maximum range. It is also assumed that if there is maximum backscatter at one frequency, there is minimum backscatter of the other frequency. The disadvantages are they have a higher probability of unwanted detection and they use more frequencies and bandwidth.

14.13.5 Chirped radar

Analog Pulse Compression or Chirped Radar uses compression techniques to improve the resolution. The requirements for a radar is to; transmit long duration pulse of high energy, detect a short duration pulse to only detect one or two radar range bins, and to optimize for both fine range resolution and long-range detection. The chirped radars use a chirp type waveform with linear frequency modulation that changes from low to high frequency during the pulse. An analog filter is used at the receive end with nonlinear phase response. The filter has a time lag that decreases with frequency. The rate of time lag decrease is matched to the rate of increase in the chirp. This results in a very short, high amplitude output from the filter. Therefore, the response of the pulse detection has been “compressed,” see Chapter 2.

14.13.6 Digital pulse compression radar

Digital Pulse Compression Radar using Binary Phase Shift Keying (BPSK) can also perform pulse compression using a matched-filter correlator detection. The transmitted radar pulse uses a pseudorandom sequence of phase modulations and this is detected by using a filter matched to that same sequence. The resulting output will match only when the stored sequence matches the received sequence. The timing resolution of the receive signal is equal to the transition time of the phase changes. The detection method can also filter out undesired signals that do not match the stored sequence and improve resolution of the received pulses, see Chapter 5.

14.13.7 Frequency-modulated CW radar

Continuous wave Radar is frequency modulated and is transmitted to the target. The echo from the target returns a delayed version of the same signal to the receiver (Figure 14.43). Multiplying the transmitter waveform with the delayed received waveform produces a beat frequency or difference frequency. This beat frequency is proportional to twice the distance of the target. If the target is moving, a Doppler frequency is produced which correlates to the velocity of the target. This radar requires good isolation between the transmitter and receiver antennas.

There are several techniques uses to generate the FM signals. The first method uses a sinewave to generate the FM of the transmitted signal. The received signal is delayed in time which is directly proportional to the range. The transmitter and receiver signals are compared and the resultant is a sinusoidal beat frequency which determines the range (Figure 14.44(a) and (b)).

Another technique uses a Sawtooth waveform in one direction to FM the frequency which is called chirped-FM (Figure 14.45).

The Sawtooth changes the slope of the modulation frequency:

Δf_{fm} is the change of modulation frequency $f_1 - f_0$
Δt_{fm} is the modulation time – time it takes to change from f_0 to f_1
Δf_{radar} is the frequency change between the transmit and receive frequency = beat frequency
$\Delta f_{fm}/\Delta t_{fm} = \text{slope}$
$\Delta f_{radar} = \Delta t_{radar} \times \text{slope} = \Delta t_{radar} \times \Delta f_{fm}/\Delta t_{fm}$

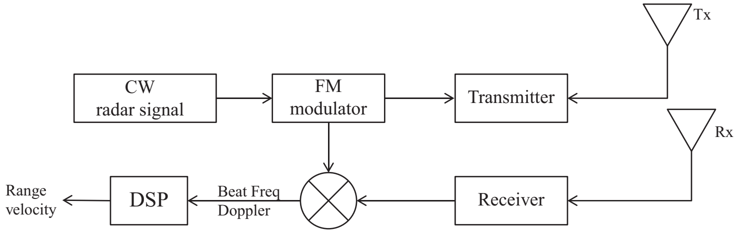


Figure 14.43 Frequency modulated CW FM-CW radar (CWFM)

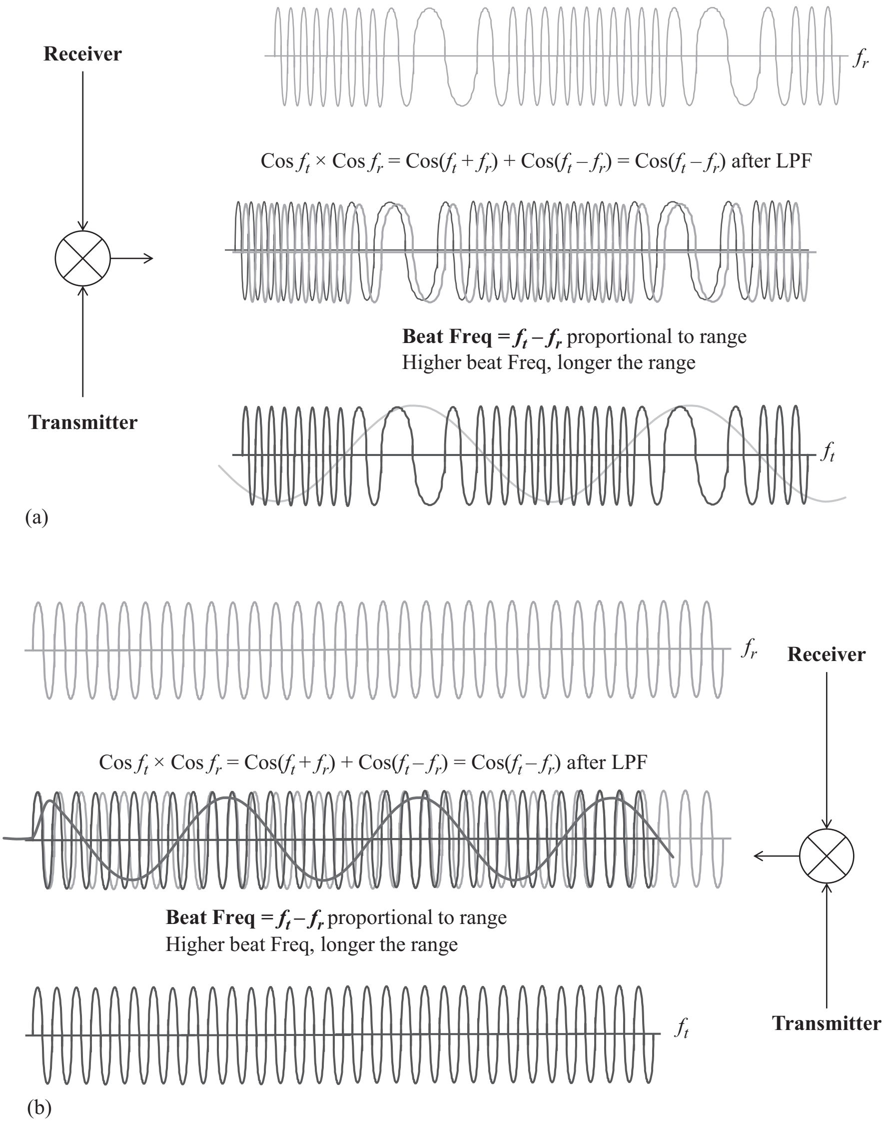


Figure 14.44 (a) Sinewave FM Transmission and (b) Sinewave FM Detection

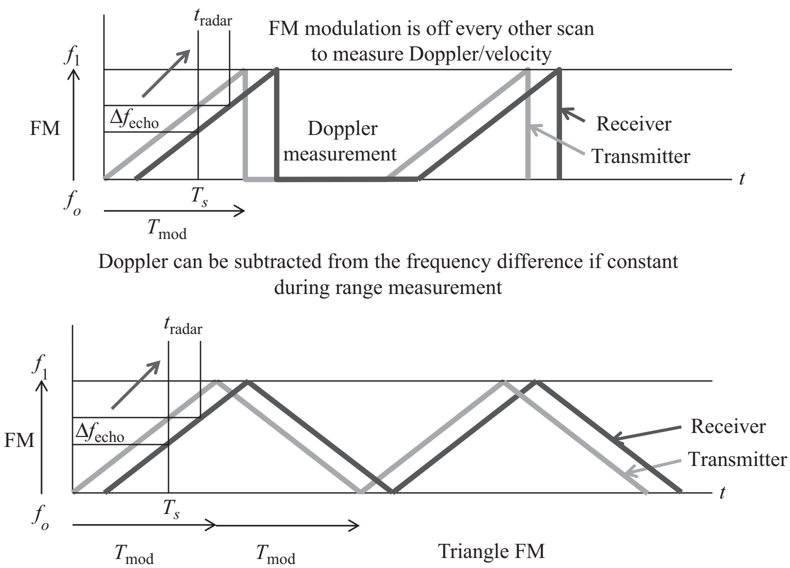


Figure 14.45 Sawtooth and triangle FM for range and velocity

$\Delta t_{radar} = \Delta f_{radar} \times \Delta t_{fm}/\Delta f_{fm}$
Range $= c_o \times \Delta t_{radar}/2$
Range $= c_o \times (\Delta f_{radar} \times \Delta t_{fm}/\Delta f_{fm})/2$

Range limit $= 0.5 \times c_o \times \Delta t_{radar}$ since receive samples are not processed for a brief period after the modulation ramp begins because incoming reflections will have modulation from the previous modulation cycle.

The sawtooth FM and a triangle FM are used for both range and velocity (Figure 14.45). The Doppler can be subtracted from the frequency difference if constant during range measurement. The triangle FM uses both up and down directions.

Another method uses a square wave, which is a two-channel frequency hop.

The range for CWFM demodulation is limited to 1/4 wavelength and the range measurements are only reliable to 60% of the instrumented range, for example, up to a range of 300 km for 100 Hz FM:

$$f_r - f_t = c_o/(4 \times f_{mod})$$

The FM sweep time is at least 5–6 times the round trip time.

The common errors in FM-CW Radars are

1. Transmitter Leakage into the Receiver—need to ensure good isolation between them. Cancelation methods: sample Tx signal, subtract from Rx signal
2. Nonlinear Frequency Sweep—Linearize the sweep, compensate the non-linearities in the detection DSP process

3. FM-to-AM Distortion—Eliminate AM, may not be required in an FM system, use AGC and saturation methods to reduce AM, compensate for the AM in the detection DSP process
4. Measurement Errors—Improve the measurement techniques, compensate for errors in the detection DSP process
5. DC Offset and $1/f$ Flicker Noise in Direct Conversion Receivers—Use a Superheterodyne receiver—double conversion

14.13.8 Doppler radar

The Doppler radar uses the Doppler Effect to measure distance and velocity of a moving target. It calculates the difference between the transmit frequency and the reflected receive frequency.

The change of frequency determines the radial velocity of target where the radial speed is in the direction from the target to the radar. For example, a target going around the radar in a circle has a Doppler frequency of zero whereas a target going toward or away from the radar produces the maximum Doppler frequency.

Doppler radars are used in weather, sea-based, semiactive homing, and astronomy radars. It produces information about target velocity during the detection process and has techniques to detect fast-moving small objects near slow-moving large objects.

The Doppler shift depends on the radar configuration whether it is active or passive. The active radar transmits a signal that is reflected back to the receiver whereas the passive radar depends upon the object sending a signal to the receiver. Doppler can also be used to separate different objects based on speed. A pulse Doppler radar signal processing uses this technique and is often combined with conical scanning or Monopulse to improve track reliability. The ability to separate different objects is necessary to separate the object signal from the interference to avoid being pulled-off the object as in pull-off jamming. The Doppler radar also avoids problems when aircraft are flying too close to the ground or through clouds.

Conical scan and Monopulse antennas are susceptible to interference from weather phenomenon and stationary objects; however, pulse Doppler radars minimize these effects. Doppler measurements require that the Nyquist sampling criteria be satisfied by sampling at twice the highest frequency in the band. An example of a Doppler weather radar is as follows:

Given: Transmit Freq $= 1$ GHz, Sample rate $= 2$ kHz, $c_o = 0.3 \times 10^9$, velocity $= 75$ m/s
Doppler frequency $= 2\ \mathrm{V}f/c_o = 2 \times 75 \times 1/.3 = 500$ Hz
Nyquist criteria $= 2 \times (500\ \mathrm{Hz}) = 1\ \mathrm{kHz} < 2$ kHz sampling rate

This example meets the Nyquist criteria so it is able to track the velocity at 75 m/s. If the velocity changes to 500 m/s, the following analysis follows:

Given: Transmit Freq $= 1$ GHz, Sample rate $= 2$ kHz, $c_o = 0.3 \times 10^9$, Velocity $= 500$ m/s
Doppler frequency $= 2\mathrm{V}f/c_o = 2 \times 500 \times 1/.3 = 3.3$ kHz

Nyquist criteria = 2 × (3.3 kHz) = 6.6 kHz < 2 kHz sampling rate
Does not meet Nyquist criteria—Does not track velocity at 500 m/s

This example does not meet Nyquist criteria so it is not able to track the velocity at 500 m/s.

14.13.9 Next-generation radar weather radar

Next-generation radar weather radar (NEXRAD) is a network of 160 high-resolution S-band Doppler weather radars. It is operated by the National Weather Service and uses the Weather Surveillance Doppler Radar WSR-88D. It detects precipitation and atmospheric movement or wind and displays a mosaic map showing patterns of precipitation & movement. National, regional, and state radar weather images are produced by compiling data from many NEXRAD sites. This radar uses two basic modes which are selectable by operator. They are clear-air mode which is slow scanning and is used when there is no significant precipitation in the area, and normal or precipitation mode which is fast scanning. The clear-air mode is very sensitive and will detect even minute echoes. It utilizes VCP31 or VCP32 and takes 10 min to produce an image with some radar sites using this mode even with light snow. It requires high sensitivity in order to detect snow showers since snow reflects less energy since it is less dense. The precipitation mode is used when there is significant precipitation and the radar automatically changes to precipitation mode when required. This mode provides higher resolution for relatively strong echoes but is less sensitive. This mode utilizes VCP21 and produces an image every 6 min. A special type of precipitation mode called "Severe Weather" mode utilizes VCP11 and produces images every 5 min, or utilizes VCP12 which produce images every 4 min. This mode is used for research or extreme weather such as hurricanes or tornadoes.

All of these products use the precipitation mode scale, even if the individual sites are in clear-air mode. This mode cleans up the mosaic and removes most of the nonprecipitation echoes. Images are further enhanced by applying anticlutter algorithms to remove nonprecipitation data. In the winter months, the algorithms are adjusted to make the mosaics more sensitive in order to detect snow showers that could otherwise go undetected because snow generally reflects much less energy than other forms of precipitation.

The type of precipitation is difficult to distinguish, for example, snow and light drizzle rain has approximately the same returns. Sophisticated algorithms are used to try and determine type of precipitation that is present. Most algorithms use current conditions data with model data as it hits the ground which is called a "Winter Mask." This mask is applied to a radar image for rain, snow, and mixed rain and snow.

Volume Coverage Patterns (VCPs) are used for each NEXRAD site to determine the amount of coverage. The site contains a 28 foot diameter dish antenna with coverage of 360° Azimuth and 20° Elevation. The precipitation usually reflects at least 15 dBz. Z is a comparison between the radar reflection levels off an object which acts as a reference to the reflection level off a rain droplet. This helps

to determine the type of precipitation present using a common reference level. The coverage search pattern scans 360° in azimuth, then increases elevation angle and scans again until the total coverage is completed. The preprogrammed set of scanning elevations is referred to as VCPs. After VCP is complete, the data is processed and images generated and displayed on the map. Some commonly used VCPs are

VCP31—Clear-Air, 10 min scan time
VCP21—Precipitation, 6 min
VCP12—Better Precipitation detection impacted by terrain blockage—4 min
VCP121—Wind velocity—5 min, also different PRF

Nonprecipitation echoes and clutter cause unwanted weak echoes that appear on radar image when there is no precipitation. The three main categories that cause this interference are; atmospheric effects, ground clutter, and false echoes.

Atmospheric effects include clouds, smoke, fog, inversion layers, and variation in air density due to temperature. Ground clutter include ground obstacles, buildings, mountains, antenna towers, and also above ground including aircraft, birds, and insects. False echoes are variations in air density that cause radar signals to refract or bend into the earth. The reflected signal is then refracted back to the dish where it shows up as a very strong echo. This occurrence is rare but is included for completeness.

NEXRAD sites send out two different sets of data; level II data which are images that use 128 colors with a resolution of 0.5 dBz and level III data which are images that use only 15 colors with only resolution of 5.0 dBz. The higher the number of colors provided the more detail of storm intensity. This allows the ability to locate certain weather phenomena such as gust fronts and hook echoes that can be hard to see with the level III data. The level II radars are high definition and are considered to be the best radar for this type of application.

14.13.10 Terminal Doppler weather radar (TDWR)

The Terminal Doppler Weather Radar (TDWR) is generally located near major airports. This radar has a resolution that is nearly twice that of the NEXRAD radars. This provides for better details on small features in the precipitation patterns. The higher resolution is only available up to 135 km from the radar. For ranges greater than 135 km, the resolution is about the same as the NEXRAD radars. The disadvantage of this radar is that it uses a shorter 5 cm wavelength (6 GHz) instead of the 10 cm wavelength (3 GHz). The shorter wavelength cannot propagate through heavy rain very far and could miss severe weather patterns. This radar should be used in conjunction with the NEXRAD radar.

14.13.11 Synthetic-aperture radar (SAR)

The SAR uses platform motion for different beam positions. The platform moves the antenna to different positions on the same target. This uses a mounted single beam forming antenna on a moving platform and monitors relative motion between

a radar antenna and the target. It provides distinctive long-term coherent-signal variations, finer spatial resolution than conventional beam-scanning means, and is used for both aircraft and spacecraft. The target scene is repeatedly illuminated with pulses of radio waves at wavelengths anywhere from a meter down to millimeters. It provides resolution imagery due to the long-range propagation characteristics and provides complex information processing capability of modern digital electronics.

The SAR receives several echoes successively at different antenna positions. These echoes are coherently detected, stored, and postprocessed in order to resolve elements in an image of the target region. Airborne systems provide resolutions to about 10 cm, whereas ultrawideband systems provide resolutions down to a few millimeters. For example, Terahertz SAR has provided submillimeter resolution. SAR images have wide applications in remote sensing and mapping of the surfaces of both the Earth and other planets. SAR can also be implemented as "inverse SAR" by observing a moving target over a substantial time with a stationary antenna.

14.14 Radar communications

Radar consists of high power, long-range equipment using burst RF transmissions to illuminate targets for detection of range and bearing. With this equipment in place, communication modulation techniques can use the radar to provide long-range communications and transfer digital data from sensors and other sources. Since the radar was designed to use high power pulses with low duty cycle with low average power, communication and data link techniques incorporate the modulate to accommodate the pulse framework.

Two basic modulation approaches have been developed that utilize radar directly, burst communication by modulating the carrier inside the pulse, and pulse coded modulation (PCM) and pulse position modulating using the position of the radar pulses for encoding/decoding data.

14.14.1 Direct burst pulse-coded modulation (PCM) for communications

This type of radar communications uses the carrier in the pulse to transmit voice and data. The technique modulates the CW carrier directly using Phase Shift Keying, Frequency Shift Keying, or other types of modulation. By modulating the carrier directly, for example, using a Binary Phase Shift Keying (BPSK) modulation, the spectrum in the frequency domain is generated (Figure 14.46). This type of modulation phase shifts the carrier depending on the data that is sent. Therefore, the data is encoded in the phase shifts of the carrier.

The radar produces the pulse train which includes the distance between the pulses or PRI, and this modulates the carrier frequency using on/off keying (Figure 14.47). The data is encoded including timing of the radar pulse and sends out bursts of data that is synchronize with the radar pulse. The encoded data modulates the radar CW pulses with the type of modulation selected. This is shown using a

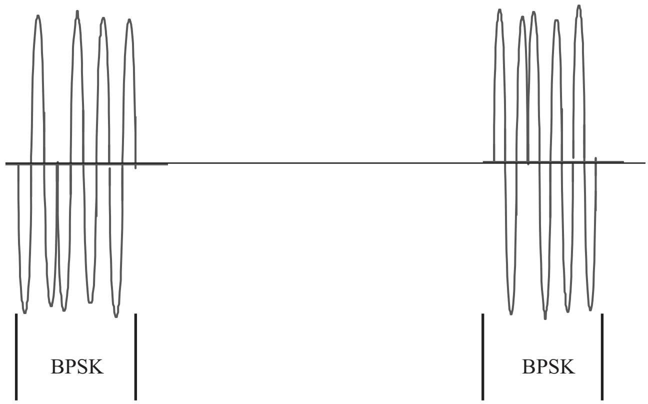


Figure 14.46 Direct modulation of the radar pulse carrier provides communications

spectrally efficient Continuous Phase-Shift Keying (CP-PSK) modulation. The output of the transmitter is a radar burst with a phase modulated carrier frequency (Figure 14.47). Since most radars operate in or near saturation, a constant envelop PSK is used to reduce sidelobe regeneration and distortion due to saturation. One of the best waveforms to use with radar is CP-PSK for several reasons. This waveform performs well in saturated applications and can change phase states using digital commands. This constant envelope, continuous phase modulation is spectrally efficient since the sidelobes are reduced. It contains no amplitude modulation which prevents spectral regrowth of the sidelobe during saturation. It can utilize a root raised cosine filter to reduce intersymbol interference and can add forward error correction such as RS, LDPC, or Turbo coding if needed.

Modulating the carrier direct is referred to as PCM. This is generally high speed modulation due to the fact that there is little time to transmit the data during the length of the radar pulse.

The receiver for radar communications accepts the incoming modulated pulses and detects and synchronizes the pulse and carrier using a carrier recovery loop. The carrier recovery loop strips off the carrier and the code recovery loop produces the data with a bit synchronizer to detect the data, see Chapter 2. Additional Spread Spectrum can be used for antijam, LPI, and LPD applications in addition to sending the data only. This modulated radar pulse is only on for a short amount of time during transmission, the waveform has even better antijam/LPI/LPD than continuous waveforms against electronic threats. For example, common burst/pulsed jammers have minimal effect on the burst unless it can determine and synchronize to the PRF. Also, many radars use varying PRFs which makes it more difficult to jam. Some other advantages include

- Minimal interference caused by other radars
- Ideal for half-duplex two-way communications

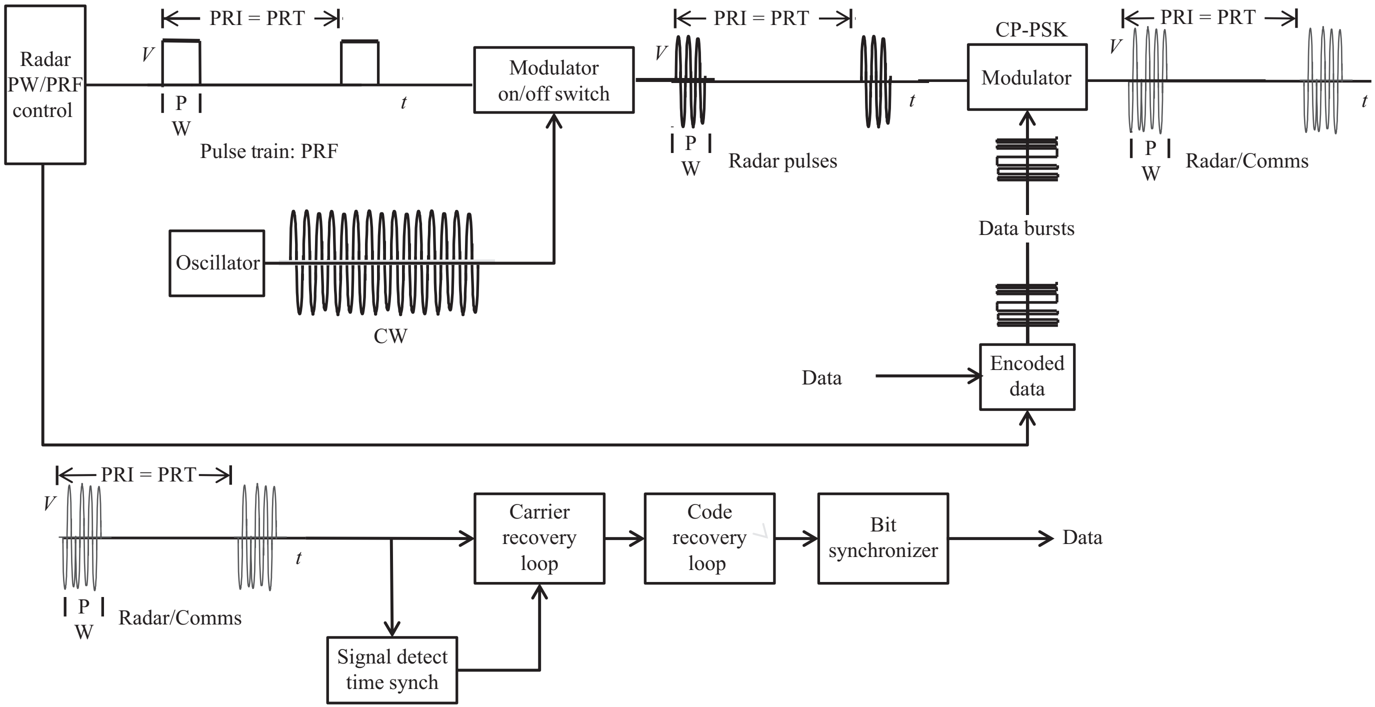


Figure 14.47 Direct modulation/demodulation of the radar pulse

- Uses TDM for networking applications
- Creates Radar Pulse Compression for Range Resolution improvement

14.14.2 Pulse-coded modulation/pulse position modulation

PCM can be combined with PPM to provide a robust, spread spectrum data link using pulsed or radar applications. PCM uses a pseudonoise PN code for spread spectrum to help mitigate jammers and provide an LPI/LPD waveform. PPM is used to encode data by detecting the pulse in time to represent data bits (Figure 14.48). The data bits represent a time slot for PPM and the spread spectrum is provided by the PCM. The modulation uses matched filter correlators to correlate the PN codes. The integrated correlated signal produces a pulse, and the time position of pulse provides the data information. The number of bits is dependent on the number of time slots used:

$$2^{\#\text{bits}} = \#\text{ of time slots}$$

For 8 time slots each pulse represents 3 data bits:

$$2^{3\text{bits}} = 8\text{ time slots}$$

Data is retrieved by the time position of the pulse using either absolute PPM or differential PPM, refer to Chapter 5 for more details.

The radar sends out the burst transmissions of PCM which provides the pulse for the PPM. The PCM waveform provides process gain using spread spectrum. An example of the process gain that can be achieved is shown below:

Given: PW = 100 μs, PN-code rate = 10 MHz
PN-code time = 1/10 MHz = 100 ns
\# of modulations/pulse = 100 μs/100 ns = 1,000
Process gain = 10log1,000 = 30 dB

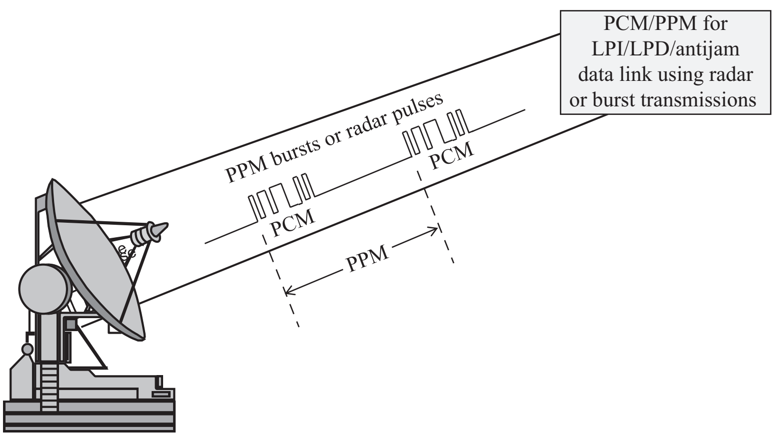


Figure 14.48 PCM/PPM communications integrated with radar

This provides good for antijam and LPI/LPD. The PCM is processed by an asynchronous match filter correlator to realize the process gain (Figure 14.49). The output of the matched filter correlator is a high-level pulse encoded with the data and the position of the pulse or PPM is decoded to retrieve the data.

The modulator is similar to the previous PCM modulator with the addition of changing the time of the pulse to encode the data (Figure 14.50). The demodulator uses the carrier recovery to strip off the carrier, and then uses matched filter correlators to remove the spread spectrum that generates the pulse which is decoded into data according to the time position of the pulse. Stages of match filter correlators can be used to increase the process gain over jamming signals and then the data is synchronized and decoded (Figure 14.50), see Chapter 2 for more information on PCM/PPM.

The time position of the pulse represents a stream of bits for digital communications and data transport (Figure 14.51). This uses the same modulated radar pulse with a time slot location representing the data to be sent. This modulation can be easily adapted and used with existing radars to provide an integrated data link. The radar is an existing infrastructure with high gain antennas, long-range capabilities, and the radar burst provides spread spectrum communications and data transfer using PCM/PPM. A block diagram of the PPM transmitter and the PPM receiver are shown in Figure 14.52. The digital data is formatted for the data structure including a root raised cosine to reduce ISI, interleaved to enhance the FEC for burst errors, and FEC is provided to correct errors if needed. The time slot grid is developed for the PPM with the external time reference from the radar to synchronize the time. This waveform is then formatted depending on the application, burst, pulse, or laser generator. This example shows 16 time slots that are used to encode the data, t0–t15 (Figure 14.52). The receiver accepts the burst, pulse, or

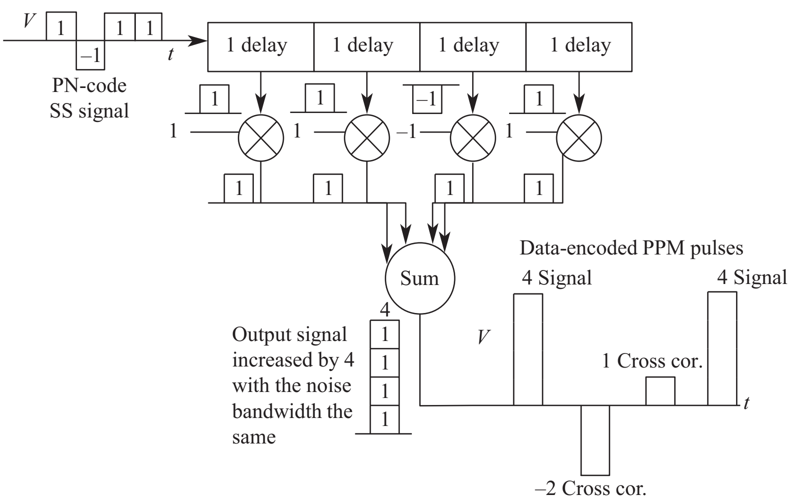


Figure 14.49 Detection process for PCM/PPM using a matched filter correlator

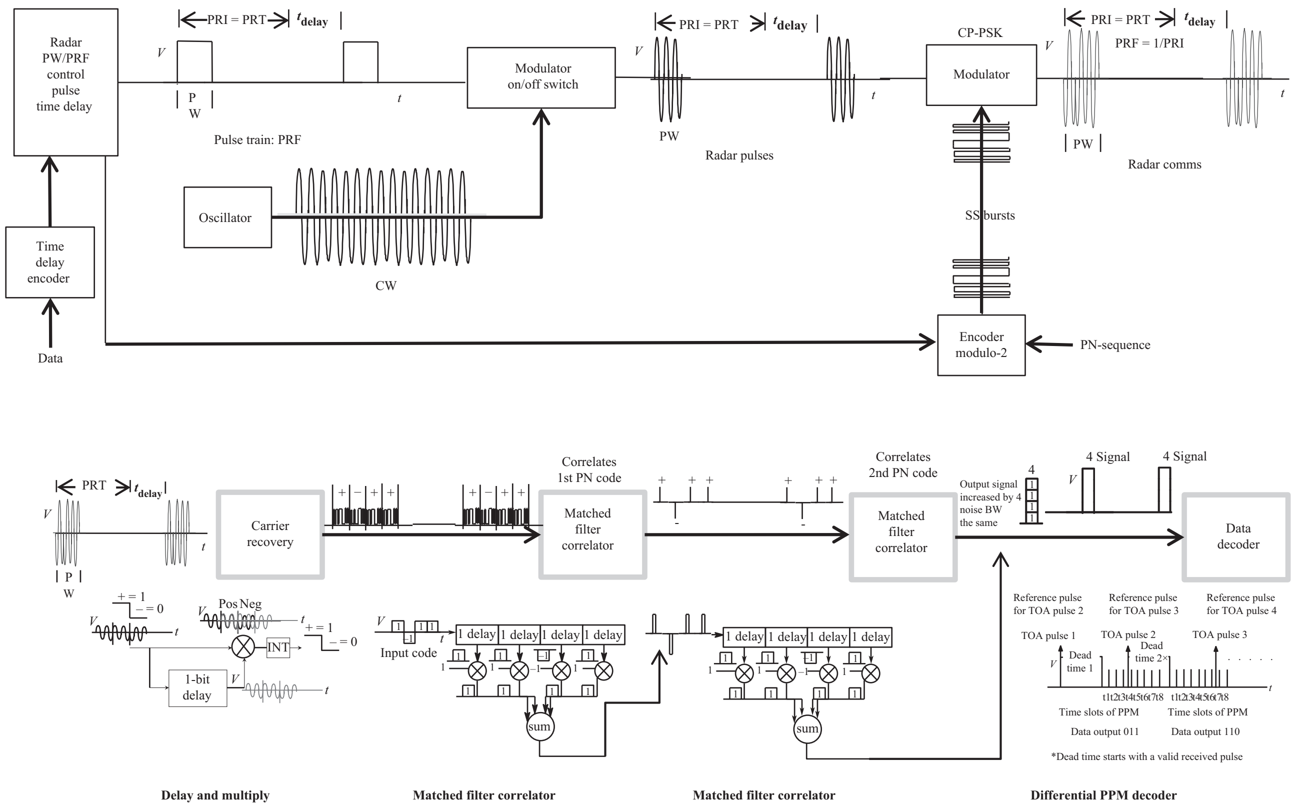


Figure 14.50 Direct sequence spread spectrum with pulse position modulation (PPM)

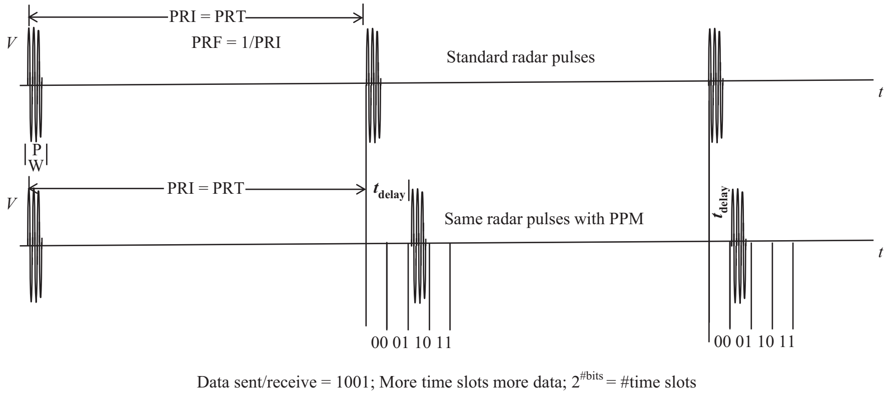


Figure 14.51 Radar data link using pulse position modulation (PPM)

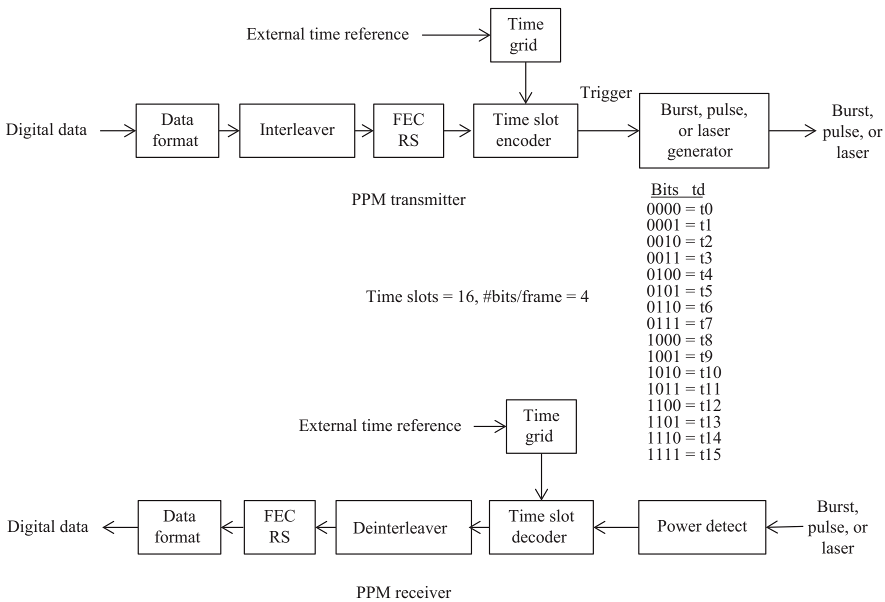


Figure 14.52 PPM system block diagram

laser communication signal and does a power detect which is used to determine what time slot it was in to decode the data.

There are many applications to radar communications including using a common data link CDL with a radar R-CDL. It uses the CDL in a burst mode instead of a continuous mode to be used with the radar system. Also, there is a Half-Duplex CDL where it is transmitting for half the time and receiving for half the time. This half-duplex system works well with total isolation between the transmitter and the

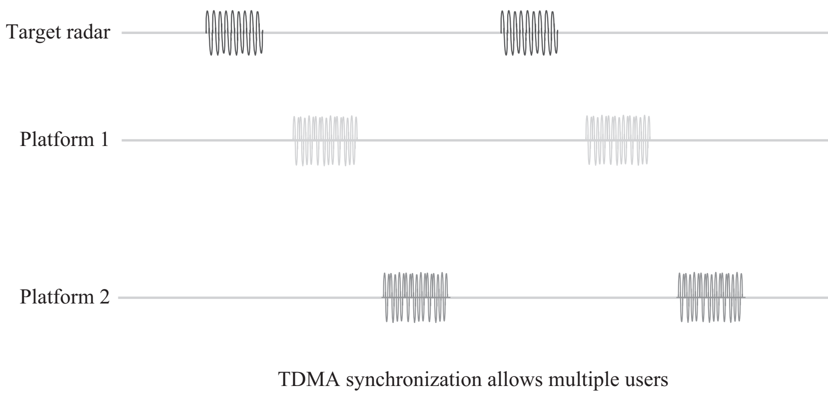


Figure 14.53 Simultaneous transmit/receive using TDMA

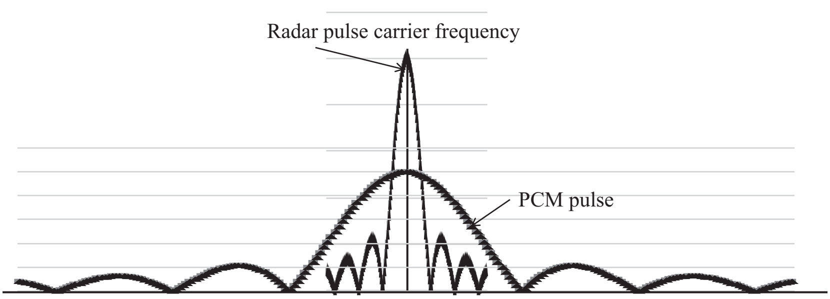


Figure 14.54 Bandwidth of a radar pulse vs PCM pulse

receiver and eliminates T/R interference in a network environment. Another application uses TDMA for Multiuser functionality (Figure 14.53).

Using PCM increases the bandwidth compared to the radar pulse carrier (Figure 14.54). This wider bandwidth provides data modulation and spread spectrum if applicable and can provide process gain to improve the system against jammers and improves the LPI/LPD. The band spreading also improves the radar by increasing the range resolution in the spreading and dispreading of the pulse compression.

For example: Given: PW = 100 μs. Modulation rate = 10 MHz

Time of modulation = 1/10 MHz = 100 ns
Number of modulations/Pulse = 100 μs/100 ns = 1,000
Process gain = 10log1,000 = 30 dB
Modulation rate at 100 MHz—Process gain = 40 dB
Uses a PN-sequence to modulate the CW radar burst—CP-PSK

14.14.3 Radar communication applications

There are many applications for using radar equipment to send communications and data. An ideal application is using a missile tracking radar to track the missile and at the same time incorporate a PCM/PPM modulation of the radar burst to send control and detection data to the missile and the missile sends back the target information back to the radar. This can be synchronized to avoid collision in the radar application and the two-way communication capabilities (Figure 14.55).

Radar communications can also be used to transmit policy-based management that needs to be distributed to multiple operational data links to receive the policy information since radar can reach long distances with maximal coverage. This can be accomplished by broadcasting the policy-based information from the radar system. Radar communications can also provide adaptive processes such as Dynamic Spectrum Allocation to the communication radios and networks to synchronize changing of operational frequencies. The radar still functions as a radar at the same time or at separate times providing both radar and communications. In addition, radars can be used for discovery waveforms for networks to find new users that want to join the network using directional, high gain, long-distance antennas. Also, radars can be used for directional MANET applications to provide control for the networks.

Radar communications can be used for satellite to satellite communications using PPM for Doppler mitigation and TDM to form a network according to a timing grid (Figure 14.56). Each of the satellites is assigned to a timing slot for communications, and the data is encoded in the PPM time grid. Spread spectrum

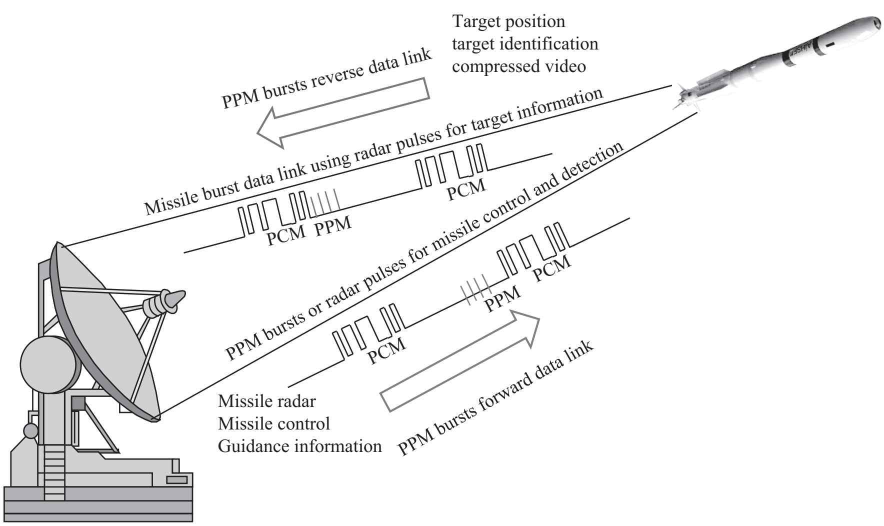


Figure 14.55 Radar use includes missile control and target data information transfer

using PCM can also be used to prevent interference and jamming. Another option in place of time separation and TDM is to separate the users in frequency or FDM (Figure 14.57). The same PPM waveform is used, but different frequencies separate the satellites in the network.

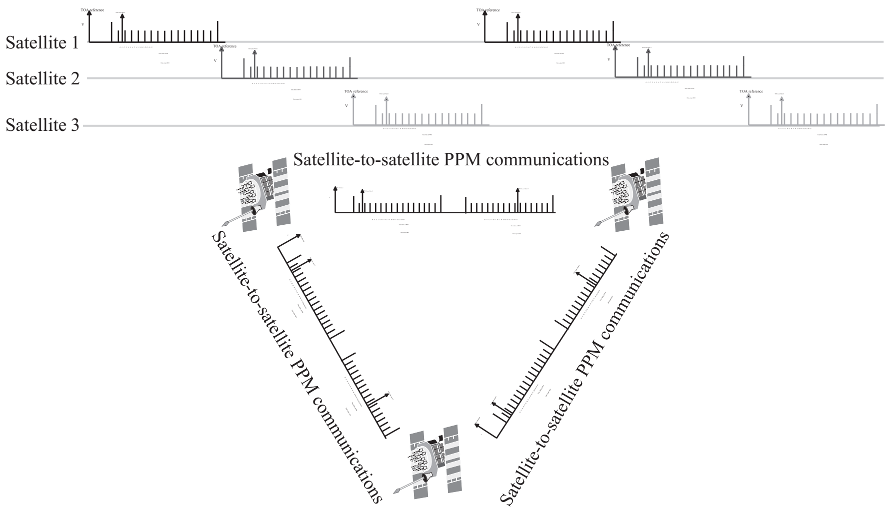


Figure 14.56 Satellite communications systems using synchronized PPM & TDM

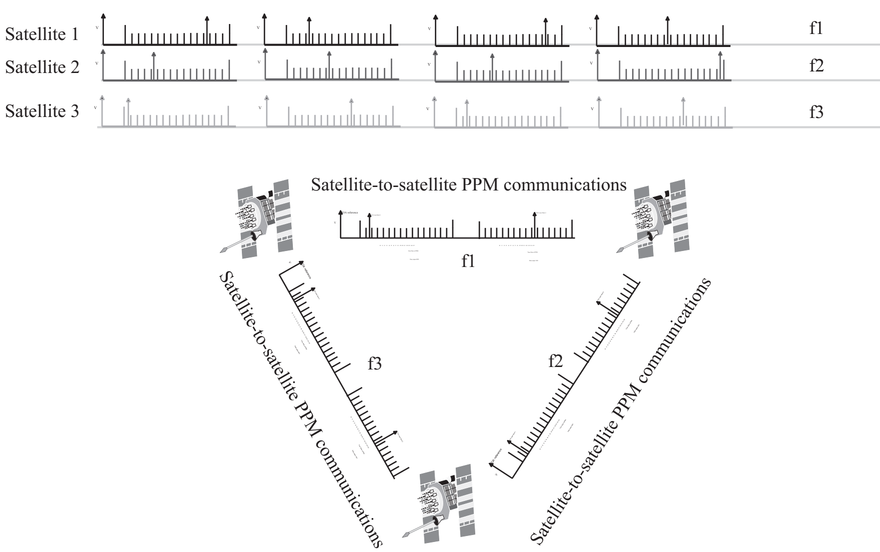


Figure 14.57 Satellite communications using PPM and frequency division multiplexing (FDM)

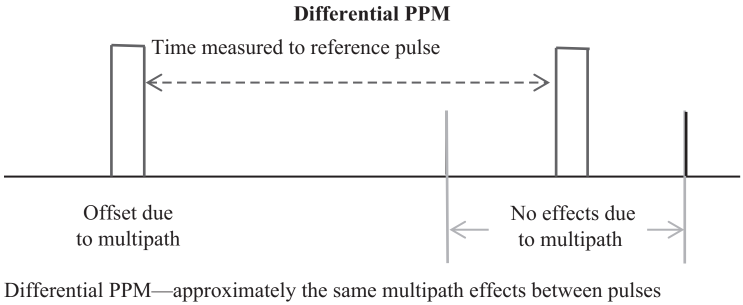


Figure 14.58 Multipath mitigation using differential PPM

14.14.4 Advantages of PCM/PPM against Doppler and multipath

The PCM/PPM system is very robust against Doppler effects since it is independent of phase/frequency. Therefore, it does not require compensation or an equalizer against Doppler. PPM is only dependent on amplitude and time of arrival instead of standard PSK modulation that is dependent on phase which the Doppler changes.

Data is retrieved even with large amount of Doppler spreading and phase changes. PCM is used for spread spectrum, antijam, LPI/LPD, multipath mitigation, encryption, and when used with PPM generally does not contain data. However, if higher data rates are required, data could be encoded/decoded with both PCM and PPM. Doppler effects cause amplitude degradation of the PPM signal which depends on the amount of Doppler but does not distort the data, just reduces the range. For high Doppler with a BER of 10^{-2} is catastrophic for PSK and FSK systems, however, with PCM/PPM since it is only dealing with amplitude and timing only degrades the amplitude by a small amount, but as long as the power level is detectable, the BER doesn't change for the PPM timing grid. With a chip error rate $= 10^{-2}$ for the PCM direct sequence spreading, the PPM waveform's amplitude exhibits approximately 3 dB loss due to degradation of PCM. Therefore, with sufficient S/N, the PPM pulse overcomes the high Doppler effects with no errors.

Using differential PPM helps to mitigate any multipath effects since offsets in timing due to multipath have approximately the same offsets between adjacent pulses (Figure 14.58). Since the second pulse timing is referenced to the first pulse, if they both shift with multipath, the time between them remains constant. Differential techniques are used for many applications to mitigate multipath, Doppler, oscillator drifts, and any changes over time that might affect the integrity of the signals. Another way to mitigate the effects of multipath is widening the time slot to compensate for the time shift of the pulse. Multipath is generally delayed and lower amplitude than the desired signal (Figure 14.59). The combination of the desired signal and multipath will adjust the timing slightly, and if the time slot is wide enough, the multipath with be mitigated. Another technique is to use leading edge tracking of the highest pulse or average pulse which keeps the frame width the same and mitigates the multipath effects. In addition, the frame width can be

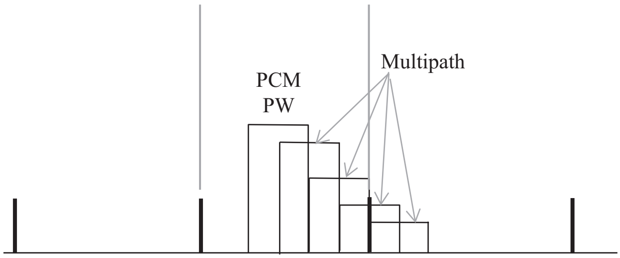


Figure 14.59 Other multipath mitigation techniques for PPM

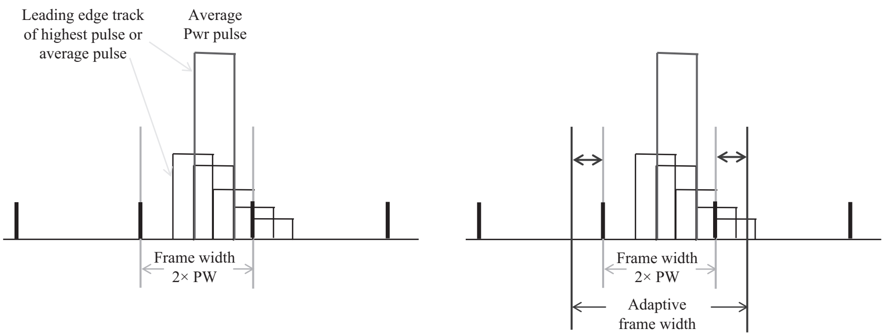


Figure 14.60 Additional multipath mitigation technique

adaptive to the system so when there is more multipath present, the frame is adjusted (Figure 14.60).

PPM is ideal for radar communications since it is adaptable for use in existing radar equipment with minimal effort.

14.15 Conclusion

Radars are used in both the military and the commercial markets and provide useful information on multiple types of targets. Radars provide target position information and have the ability to discover, acquire, and track these targets. The basic radar equations are derived by examining the signal and noise levels through the system using a link budget analysis. There are many applications for radar including using radars to provide wireless communications and data transfer. Radar equipment provides high gain antennas, long-range coverage, and is becoming more versatile incorporating AESA technology. With the new mobile devices, there are many applications that are using information from radars to provide weather details to the end user. Radars are being used for communications, transmitting sensor data, and sending out control information for missile control and others. Modulation techniques are being developed to prevent problems in saturation using CP-PSK PCM

and using PPM to reduce Doppler and other effects with minimal changes to the radar system.

14.16 Problems

1. What does RADAR stand for? When did RADAR start to be used?
2. What are the two basic types of RADAR?
3. Name one advantage and disadvantage of a pulse radar compared to CW radar?
4. How does a short PRI increase distortion?
5. What does PRI and PRF stand for and what is the relationship of PRI to PRF?
6. Define duty cycle. Given PW = 1 μs, PRF = 100 kHz: Duty Cycle = ?
7. Describe what the spectrum would look like for a 25% duty cycle?
8. What type of modulation is used for pulse radar?
9. Determine range from a reflector using sound waves given 5.38 s from voice to reflector to ear assuming speed of sound is 1,125 ft/s? How many miles?
10. What is "s" with respect to a Radar target?
11. Why is atmospheric losses times by 2?
12. What is the radar received power in dBm given the following: $P_t = 60$ dBm, $G_r = G_t = 40$ dB, $A_{fs} = -100$ dB, $G_{\text{targ}} = 21$ dB, $L_s = 10$ dB?
13. If the range is doubled, how much extra power is needed from the transmitter?
14. How fast does a radar pulse travel?
15. How do you calculate the range from the radar pulse? If the two-way time delay equals 2 ms, how far away is the object?
16. Define what is meant by range ambiguity? How can it be mitigated?
17. What determines the minimum detectable range for a radar?
18. What is meant by range resolution?
19. What methods can be used to increase range resolution?
20. What is bearing?
21. What is the difference between linear velocity and angular velocity?
22. What does PPI stand for? What does it display?
23. What is an A-scope and what does it display?
24. Describe the basic operation of a radar.
25. What does AESA stand for and what are the advantages of the AESA?
26. What is the difference between monostatic and bistatic radars?
27. What is a frequency diversity radar? What are the advantages/disadvantages?
28. Describe how a Monopulse radar works?
29. Why is a Monopulse better than a CONSCAN?
30. What is an analog pulse compression radar sometimes called? Why? How does it mitigate jammers?
31. What is clutter? How can it be mitigated?
32. What is probability of false alarm? Probability of Detect? Where is a rule-of-thumb threshold?

33. What is MTI and what principle does it use?
34. What is meant by blind speeds and what causes them?
35. What technique is used to eliminate blind speeds?
36. What modulation waveform is ideal for radar communications?
37. What is Doppler radar used for?
38. What is a SAR radar and what is it used for?
39. Given: $P_t = 1$ W, frequency $= 2.4$ GHz, $R = 100$ m, calculate freespace loss in dBm. What is the received power in dBm due to freespace loss? Do you subtract or add your answer to the power level?
40. What is the PRF given PRI $= 2$ μs? What is the carrier frequency if there are 5 cycles in a PW $= 100$ ns? What is the duty cycle?
41. What is the power level using the radar equation received at the radar given the following:

$$P_r = \frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 R^4 L_s}$$

$P_t = 100$ mW, $G_t = G_r = 316$ $f = 2.4$ GHz, $\sigma = 110$ m^2, $R = 100$ m, $L_s = 10$?
42. What is the power level in dBm received at the radar given the following: $P_t = 20$ dBm, $G_t = G_r = 25$ dB $f = 2.4$ GHz, $\sigma = 110$ m^2, $R = 100$ m, $L_s =$ 10 dB?
43. What is the maximum unambiguous range with a PRI $= 15$ μs, pulse width time $T = 100$ ns?
44. What is the range resolution given the transmitters pulse width $= 1$ μs or bandwidth of 1 MHz?
45. What is the minimum angular resolution as a distance between two target given the slant range $R = 100$ m and the 3 dB beamwidth $= 3°$?
46. What is the beamwidth of a parabolic antenna with a diameter of 3 m operating at 2.4 GHz?
47. What is the maximum angular velocity in radians/second of a target going 10 m/s perpendicular to the radar with a radius of 20 m? What is the angular velocity in degrees/second?
48. What is the two-way Doppler frequency for a target moving toward the radar with a radial velocity of 70 km/h at a frequency of 2.4 GHz?
49. What is the radial velocity if the actual velocity is equal to 10 m/s and the angle of the velocity to the radar is equal to 30°?
50. What is the first blind speed given a PRF $= 100$ Hz and a frequency of 2.4 GHz?
51. Given: $f_{\text{transmit}} = 1$ GHz, sample rate $= 2$ kHz, $c_o = 3 \times 10^8$, $v_r = 75$ m/s. Calculate the Doppler frequency. What is the Nyquist Criteria required? Does it meet or exceed the Nyquist criteria?
52. Given: $f_{\text{transmit}} = 1$ GHz, Sample rate $= 2$ kHz, $c_o = 3 \times 10^8$, $v_r = 500$ m/s, Doppler frequency $= 2v_r/c_o \times f_{\text{transmit}} = (2 \times 500/3 \times 10^8) \times 1 \times 10^9 = 3.3$ kHz, what is the Nyquist Criteria required? Does it meet or exceed the Nyquist criteria?

53. Given the parameters of problem 14:
    (a) Can it track a target going at 160 m/s?
    (b) Can it track the target if 160 m/s is 45° to boresite?

Further reading

Chang K. *RF and Microwave Wireless Systems*. Hoboken, NJ: John Wiley and Sons, 2000, page 198, ISBN: 978-0471351993.

Hovanessian S. A. *Radar System Design and Analysis*. Norwood, MA: Artech House 1984.

Kingsley S. and Quegan S. *Understanding Radar Systems*. Edison, NJ: SciTech Publishing, Inc. 1999.

Nathanson F. E. *Radar Design Principles*, 2nd ed. New York City, NY: McGraw-Hill Book Company 1999.

Chapter 15
Direction finding and interferometer analysis

Direction finding is a method to determine the direction of a transmitted signal by using two antennas and measuring the phase difference between the antennas (Figure 15.1). This process is called interferometry. In addition to using a static interferometer, further analysis needs to be done to calculate the direction when the interferometer baseline is dynamic; that is, the interferometer is moving and rotating in a three-dimensional plane. Thus, coordinate conversion processes need to be applied to the nonstabilized antenna baseline to provide accurate measurement of the direction in a three-dimensional plane.

15.1 Interferometer analysis

For a nonstabilized antenna baseline, roll, pitch, and yaw of the antenna baseline need to be included as parts of the interferometer process for accurate azimuth determination using interferometer techniques. Therefore, the azimuth angle calculation involves a three-dimensional solution using coordinate conversions and direction cosines. The overall concept is that the coordinates are moved due to the movement of the structure. The direction phasor is calculated on the moved coordinates, and then a coordinate conversion is done to put the phasor on the absolute coordinates. This will become apparent in the following discussions. A brief discussion on direction cosines follows since they are the basis for the final solution.

15.2 Direction cosines

Direction cosines provide a means of defining the direction for a given phasor in a three-dimensional plane. They are the key in defining and calculating the interferometer equations.

A phasor A is represented as:

$$A = |A|(\cos \alpha i + \cos \beta j + \cos \gamma k)$$

where i, j, and k are unit vectors. For example, $i = (1,0,0)$, $j = (0,1,0)$, and $k = (0,0,1)$ related to the magnitudes in x, y, and z, respectively. For example, i is just for the x direction. Therefore, α is the angle between phasor A and the x-axis, β is the angle between phasor A and the y-axis, and γ is the angle between phasor A and the z-axis (Figure 15.2).

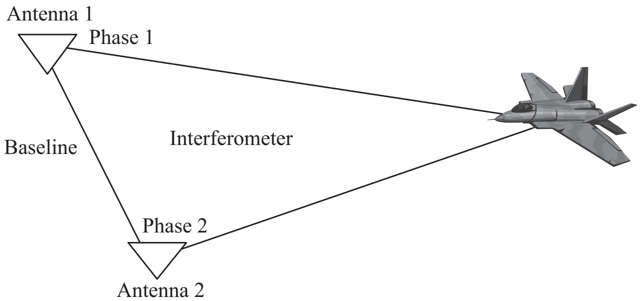


Figure 15.1 Basic interferometer used for direction finding

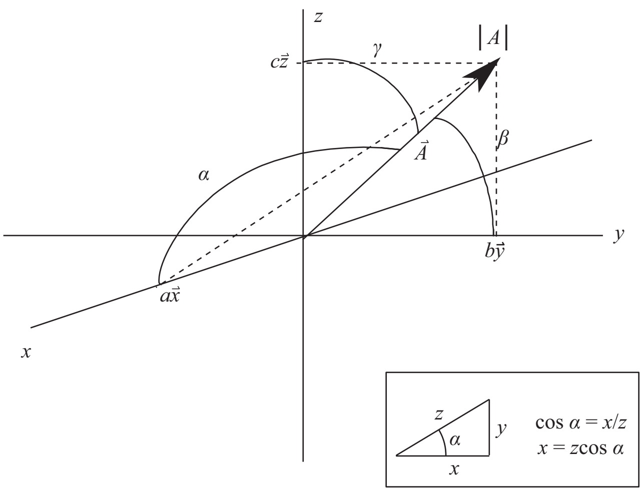


Figure 15.2 Angles for computing the direction cosines

The direction cosines are the cosines of each of the angles specified: $\cos\alpha$, $\cos\beta$, and $\cos\gamma$. The cosines are equal to the adjacent side, which is the projection on the specified axis divided by the magnitude of phasor A. The projection on the x-axis, for example, is simply the x component of A. It is sometimes written as $\text{comp}_i A$. The vector A can be described by (x,y,z). The direction cosines are defined in the same manner, with x, y, and z replacing i, j, and k.

Another phasor is defined and is the horizontal baseline in the earth's plane. This is the baseline vector between two interferometer antennas. Therefore,

$$B = |B|(\cos\alpha x + \cos\beta y + \cos\gamma z)$$

If $\gamma = 90°$, which means that the phasor is horizontal with no vertical component, then

$$A = |A|(\cos\alpha x + \cos\beta y + 0)$$

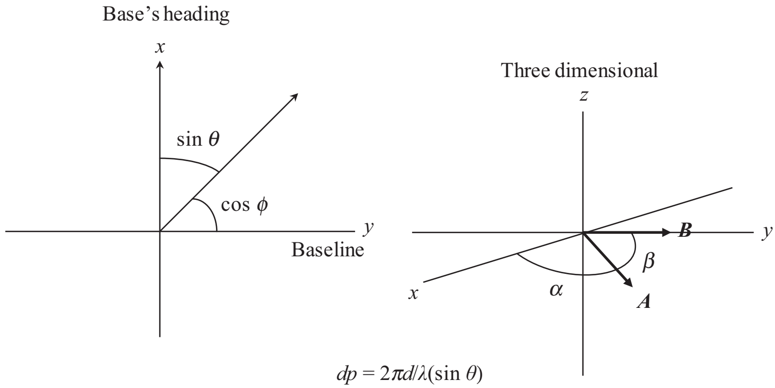


Figure 15.3 Standard interferometer two-dimensional equation with baseline on the y-axis

If the baseline is on the y-axis, then:

$$B = |B|(\cos 90x + \cos 0y + \cos 90z) = |B|(0x + 1y + 0z)$$

Therefore,

$$A \cdot B = |A|\ |B|\cos\beta$$

For this example, if the baseline interferometer is mounted on the y-axis, then α is the angle from the boresight, or the x-axis, to the baseline, which is the desired angle. This leads to the standard interferometer two-dimensional equation. For a two-dimensional case, this dot product equals $|A|\ |B|\sin\alpha$ (Figure 15.3).

15.3 Basic interferometer equation

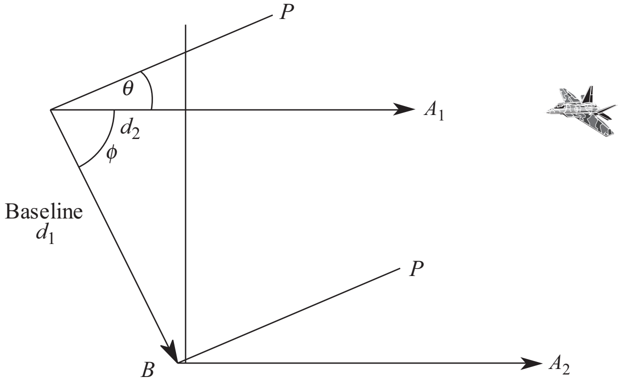
The basic interferometer equation, used by many for approximate calculations, is only a two-dimensional solution:

$$dp = (2\pi d)/\lambda \sin\theta$$

where dp is the measured electrical phase difference (in radians), d is the separation of the interferometer antennas, θ is the true azimuth angle, and λ is the wavelength. Note: θ is equal to α in the previous example.

This is a familiar equation used in most textbooks and in the direction finding literature. It is derived by simply taking the dot product of the direction phasor and the interferometer baseline, which produces the cosine of the angle from the baseline. Using simple geometry, the relationships between the additional distance traveled for one interferometer element (d_2) with respect to the baseline difference (d_1) and the measured electrical phase difference (d_p) are easily calculated:

$$d_2 = d_1 \cos(\varphi)$$



A dot $B = ab\cos(\phi)$
Opposite/adjacent $d_2/d_1 = \cos(\phi)$, $d_2 = d_1\cos(\phi)$
$dp = (d_2/\lambda)2\pi = (2\pi/\lambda)d_2 = (2\pi/\lambda)d_1\cos(\phi) = (2\pi d_1/\lambda)\cos(\phi)$
$\cos(\phi) = \sin(\theta)$
Therefore:
$d_p = (2\pi d_1/\lambda)\sin(\theta)$ = Standard two-dimensional interferometer equation

Figure 15.4 Two-dimensional interferometer

where φ is the angle of the phasor from the baseline and

$$d_p = (d_2/\lambda)2\pi$$

where d_2 is the extra distance traveled from the target.

The analysis with the baseline on the y-axis is shown in Figure 15.3.

Since the angle is usually specified from the boresight, which is perpendicular to the interferometer baseline, the equation uses the sine of the angle between the boresight and the phasor (θ), which results in the standard interferometer equation (Figure 15.3):

$$d_p = (2\pi d_1)/\lambda \sin\theta$$

This still assumes that the target is a long distance from the interferometer with respect to the distance between the interferometer elements, which is a good assumption in most cases.

This works fine for a two-dimensional analysis and gives an accurate azimuth angle with slight error for close-in targets, since $d\sin(\theta)$ is a geometrical estimate and A1 and A2 are assumed parallel (Figure 15.4).

However, a two-dimensional solution does not work for a three-dimensional solution. As the elevation increases for the three-dimensional case, the azimuth angle gets smaller and at the worst case of 90° angle the azimuth goes to zero (Figure 15.5).

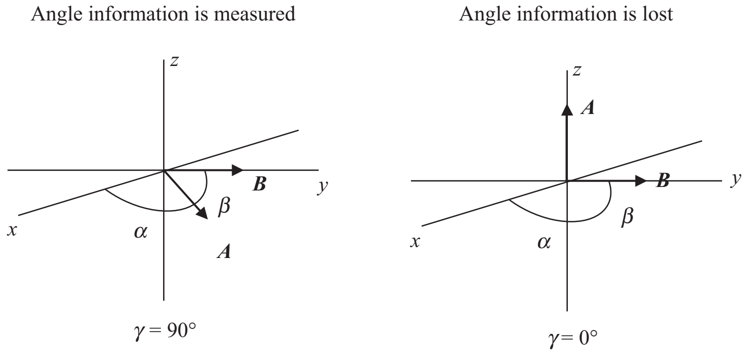


Figure 15.5 Angle information is lost with two-dimensional solution

15.4 Three-dimensional approach

If γ is not equal to 90°, then:

$$A = |A|(\cos\alpha x + \cos\beta y + \cos\gamma z)$$
$$B = |B|(0x + 1y + 0z)$$

The dot product remains the same, $A \cdot B = |A|\,|B| \cos\beta$; however, both the angles α and β change and the dot product no longer equals $|A|\,|B| \sin\alpha$. For example, if γ is 0, α and β would be 90°. Therefore, for a given A with constant amplitude, the α and β change with γ.

If the alignment is off or there is pitch and yaw, these angles change. Since the angle offsets are from the mounted baseline, then the B vector is defined with the angles offset from the B baseline (Figure 15.6). Therefore, the resultant equation is

$$B = |B|(\sin\alpha_1 x + \cos\beta_1 y + \sin\gamma_1 z)$$

where $\alpha_1, \beta_1, \gamma_1$ are the angles from the B desired baseline caused by misalignment or movement.

Therefore, if the interferometer baseline is rolled, pitched, and yawed, then the results are:

$$A = |A|(\cos\alpha x + \cos\beta y + \cos\gamma z)$$
$$B = |B|(\sin\alpha_1 x + \cos\beta_1 y + \cos\gamma_1 z)$$

This B is the new phasor, offset from the earth's coordinate system:

$$A \cdot B = |A|\,|B|(\cos\alpha \sin\alpha_1 x + \cos\beta \cos\beta_1 y + \cos\gamma \sin\gamma_1 z)$$

The main solution to this problem is to find the angle from the boresight, which is a three-dimensional problem when considering the elevation angle and the dynamics of the baseline. Approaching the problem in three dimensions using direction cosines produces a very straightforward solution, and the results are

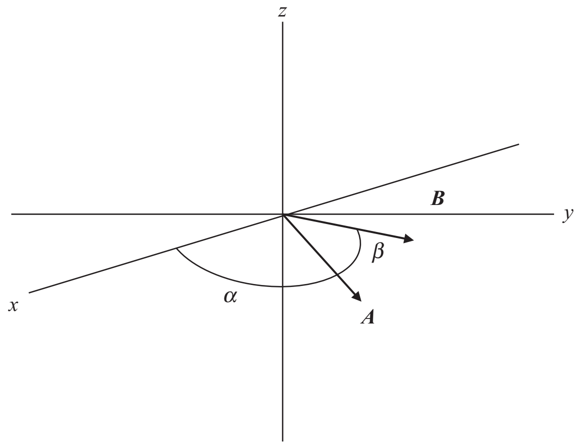


Figure 15.6 Interferometer analysis with the baseline not on the y-axis

shown in the following analysis. The standard interferometer equation is not used, and $\sin(\theta)$ is meaningless for this analysis since a two-dimensional dot product to resolve azimuth is not used. The effects that the elevation angle has on the azimuth solution are shown in Appendix C.

15.5 Antenna position matrix

The first part of the analysis is to define the three-dimensional interferometer coordinate system relative to the dynamic coordinate system. The definition of this axis is imperative for any analysis. For example, mounting an interferometer on a ship produces the following analysis using the y-coordinate from port to starboard (starboard is positive y) with the interferometer baseline along the y-axis, the x-coordinate from bow to stern (bow is positive x), and the z-coordinate up and down (up is positive z) (Figure 15.7). The position matrix for the interferometer on the ship's coordinate system is defined as $\{\sin(\alpha),\cos(\beta),\sin(\gamma)\}$.

If there is a misalignment in the x–y plane and not in the z-plane, the position matrix is $\{\sin(\alpha),\cos(\beta),0\}$, as shown in Figure 15.8. If there are no offsets, then the position matrix is (0,1,0), which means that the interferometer is mounted along the y-axis. The antenna position matrix compensates for the misalignment of the interferometer baseline with respect to the ship's coordinate system.

15.6 Coordinate conversion due to pitch and roll

The coordinate conversion transformations modify the antenna position matrix to obtain the earth coordinate antenna position x, y, and z. This is done for yaw, pitch, and roll. Heave is generally insignificant but can be included for specific cases. Yaw is the movement in the horizontal plane that affects bearing, with positive yaw

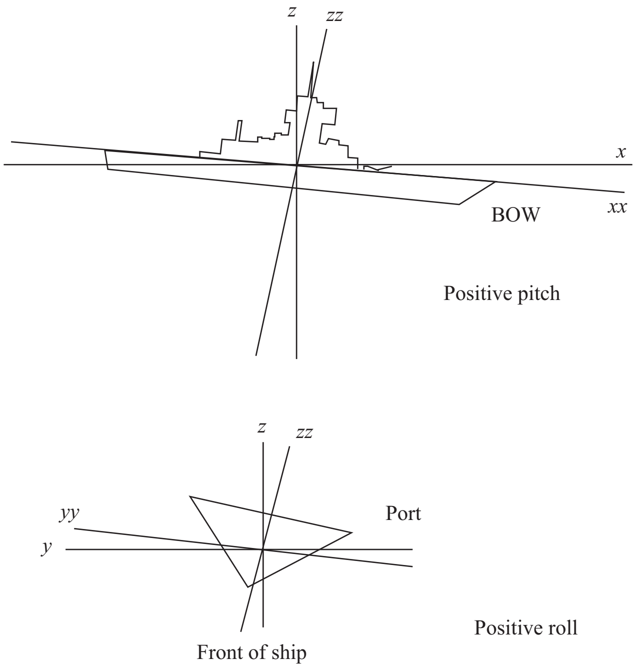


Figure 15.7 Roll and pitch definitions

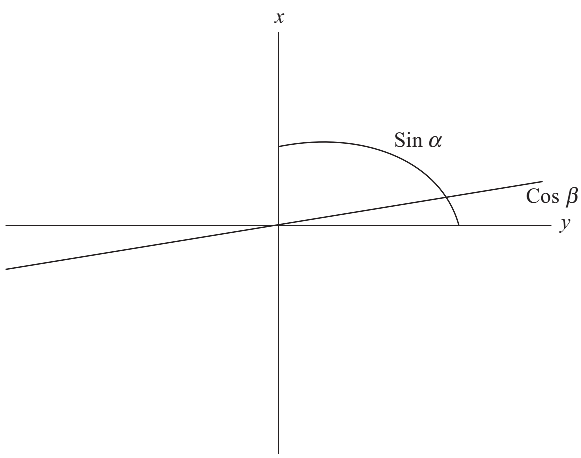


Figure 15.8 Position matrix offset

rotating in the clockwise direction. Pitch is the bow of the ship moving up and down, with down being in the positive direction. Roll is port and starboard movement of the ship rotating up and down, with port side down being positive. Heave is the movement of the ship as an entire unit in the vertical direction, with positive

$$\begin{vmatrix} \multicolumn{3}{c}{\textbf{Roll matrix}} \\ 1 & 0 & 0 \\ 0 & \cos(r) & \sin(r) \\ 0 & -\sin(r) & \cos(r) \end{vmatrix} = \begin{vmatrix} \multicolumn{3}{c}{\textbf{Roll matrix}} \\ 1 & 0 & 0 \\ 0 & 0.966 & -0.259 \\ 0 & 0.259 & 0.966 \end{vmatrix}$$

$$\begin{vmatrix} \multicolumn{3}{c}{\textbf{Pitch matrix}} \\ \cos(p) & 0 & -\sin(p) \\ 0 & 1 & 0 \\ \sin(p) & 0 & \cos(p) \end{vmatrix} = \begin{vmatrix} \multicolumn{3}{c}{\textbf{Pitch matrix}} \\ 0.996 & 0 & 0.087 \\ 0 & 1 & 0 \\ -0.087 & 0 & 0.996 \end{vmatrix}$$

$$\begin{vmatrix} \multicolumn{3}{c}{\textbf{Yaw matrix}} \\ \cos(y) & \sin(y) & 0 \\ -\sin(y) & \cos(y) & 0 \\ 0 & 0 & 1 \end{vmatrix} = \begin{vmatrix} \multicolumn{3}{c}{\textbf{Yaw matrix}} \\ 0.999 & -0.052 & 0 \\ 0.052 & 0.999 & 0 \\ 0 & 0 & 1 \end{vmatrix}$$

Figure 15.9 Coordinate transformation

heave being up. The conversion matrices are shown in Figure 15.9. This example is done for 15° roll, 5° pitch, and 3° yaw. The order of the analysis is yaw, pitch, and roll. The order of the transformation needs to be specified, and different orders will alter the final solution. Further information on coordinate transformations is included in Appendix A.

The interferometer measurement is done with the ship's antenna position at the location caused by the yaw, pitch, and roll. To bring the coordinates back to earth or level coordinates, the process is done in the reverse order and opposite direction, that is, −roll, −pitch, −yaw. The movement defined as positive pitch is bow down, and positive roll is port down. The negative numbers are put in the original coordinate transformations as shown in Figure 15.10.

Moreover, yaw can be solved after taking the roll and pitch azimuth calculations and simply offsetting the angle by the amount of the yaw. This may provide a simpler solution for implementation.

15.7 Using direction cosines

Now that the coordinate conversion and position matrix have been solved (x, y, and z values) as a result of these processes, they are then used with the direction cosines to achieve the solution. The basic direction cosine equation, as mentioned earlier, is

$$a = |a|(\cos\alpha x + \cos\beta y + \cos\gamma z)$$

For a unit vector, the scalar components are the direction cosines:

$$u = \cos\alpha x + \cos\beta y + \cos\gamma z = Xx + Yy + Zz$$

The x, y, and z values are the coordinate converted and offset compensated x, y, and z. The X, Y, and Z values are the direction cosines of the Poynting vector for the horizontal baseline coordinate system. The unit vector is equal to the phase interferometer difference divided by the phase gain.

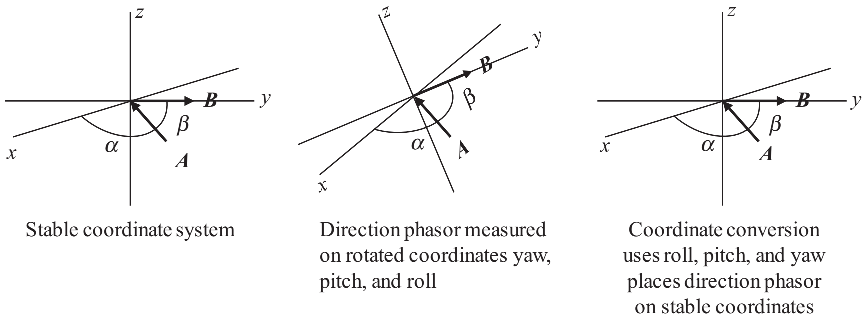


Figure 15.10 Coordinate conversions applied to nonstabilized antenna baseline

Since x, y, and z are known from previous analysis and u is measured and calculated. The direction cosines are the only unknowns and are solved by the following identities and equations:

$$\left(X^2 + Y^2 + Z^2\right)^{1/2} = 1, \text{ so } X^2 + Y^2 + Z^2 = 1$$

Note that $Z = \cos\gamma = \sin\Psi$, since the direction cosine is defined from the top down and Ψ is the angle from the horizontal up. The elevation angle (Ψ) is calculated using altitude and range. Moreover, the effects of the earth's surface can cause the elevation angle to have some slight amount of error. This can be compensated for in the calculation of the elevation angle (Appendix D). Therefore,

$$X^2 + Y^2 + \sin^2\Psi = 1$$

and from above

$$u = Xx + Yy + \sin\Psi z$$

Solving for Y

$$Y = (u - Xx - \sin\Psi z)/y$$

Solving for X^2

$$X^2 = 1 - \sin^2\Psi - [(u - Xx - \sin\Psi z)/y]^2$$

Solving simultaneous equations produces a quadratic equation for X in the form $AX^2 + BX + C$, where

$$\left[1 + (x/y)^2\right]X^2 - [2x/y(u - \sin\Psi z)/y]X + \sin^2\Psi - 1 + [(u - \sin\Psi z)/y]^2 = 0$$

where

$$A = [1 + (x/y)^2]$$
$$B = [2x/y(u - \sin\Psi z)/y] = [-2x/y^2(u - z\sin(\Psi))]$$

u = electrical phase difference (in radians) (phase interferometer difference divided by the phase gain)

$C = \sin^2 \Psi - 1 + [(u - z\sin \Psi)/y]^2$

Therefore, solving for X using the quadratic equation:

$$X = \frac{-B + \sqrt{B^2 - 4AC}}{2A}$$

and solving for Y

$$Y = \frac{u - xX - z\sin \Psi}{y}$$

then, the azimuth angle θ is

$$\theta = a\tan(Y/X)$$

The azimuth angle is the position angle from boresite (Figure 15.11).

The direction cosines are the values in the x- and y-directions in the final converted coordinate system, $Xx + Yy + Zz$. The x, y, and z coordinates include the coordinate conversion constants.

An example showing the steps in producing the interferometer solution is found in Table 15.1. This is programed into a spreadsheet for ease in changing parameters for different interferometer configurations. Other factors considered when determining azimuth are true north calculations and phase ambiguities, which are included in Appendix B. The actual excel spread sheet contains all of the equations for the interferometer example (Table 15.2).

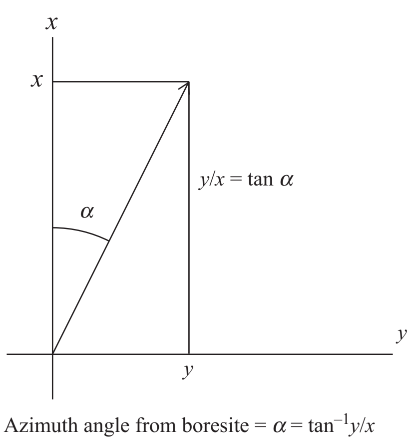


Figure 15.11 Baseline located on the y-axis

Table 15.1 Interferometer analysis excel example

Example : 15,5,3° roll, pitch, yaw, 45° EL at 24° AZ											
x-Axis on the bow–stern axis					Start with baseline						
y-Axis on port–starboard axis				Roll=	–15			**Roll and pitch only:**			
								A	B	C	
				Roll matrix		Pos. Mat.		1.0005454	–0.012517	– –0.42818	
$x' =$	0	$x' =$	1	0	0	0	x''				
$y' =$	0.9659258	$y' =$	0	0.9659258	–0.258819	1	y''	X	Y	Az. angle	
$z' =$	0.258819	$z' =$	0	0.258819	0.9659258	0	z''	0.6604613	0.2525685	20.927463	
										ATAN(Y/X)	
				Pitch =	-5						
				Pitch matrix				**Roll, pitch, yaw**			
$x =$	0.0225576	$x =$	0.9961947	0	0.0871557	0	x'				
$y =$	0.9659258	$y =$	0	1	0	0.9659258	y'	A	B	C	
$z =$	0.2578342	$z =$	–0.087156	0	0.9961947	0.258819	z'	1.0008421	0.015556	–0.428159	
				Yaw =	–3			X	Y	Az. angle	
								0.6463378	0.2867882	23.927463	
				Yaw matrix							
	–0.028026	xyaw =	0.9986295	–0.052336	0	0.0225576	x	Note that the yaw is simply			
	0.9657826	yyaw=	0.052336	0.9986295	0	0.9659258	y	an offset of 3 degrees			
	0.2578342	zyaw=	0	0	1	0.2578342	z				
				Radians	**Azimuth:**	**Answer:**		**Error:**			
Elevation angle (deg.) =			45	0.7853982	24	23.927463		0.0725368			
Electrical phase diff. (deg.) =			635.295	25.277585	Elect. phase/phase gain						
Electrical phase diff. (rad) =			11.087989	38.602861							
Phase gain =			25.132741		0.4411771	Elect. phase diff. (rad)/phase gain					
For an error in alignment, so that the interferometer is not lined up with the axis,											
the following example is shown:											
Alignment error:			**Inputs:**	**Calculations:**							
		$Y =$	0	Angle =	0						
		Dist.	1.312	$x'' =$	1						
				$y'' =$	0						

Table 15.2 Interferometer analysis example showing equations

Example: 45 degrees El, 24 degrees Az										
Definition of axis:		**Order: yaw, pitch, roll**								
x-axis on bow/stern		**Yaw = 3°, pitch = 5°, roll = 15°**								
y-axis on port/starboard		Reverse order to bring it back to the ships stabilized position: –roll, –pitch, –yaw								
Positive pitch–port down		Roll = –15°, pitch = –5°, yaw = –3°,								
Positive roll–port down		Position matrix = 0, 1, 0: y-axis on port/starboard								
						Baseline on y-axis (no offsets)				
Matrix multiply			Roll= –15			Pos matrix.		**Roll and pitch only:**		
D9×H9+E9×H10+F9×H11	= x' =	1	0	0		0	= x''	A	B	C
D10×H9+E10×H10+F10×H11	= y' =	0	Cos(–15×π/180)	Sin(–15×π/180)	×	1	= y''	1+(B14/B15)^2	(–)2×B14/B15^2×(D28/D29–B16×(SIN(E26)))	(SIN(E26))^2–1+((F29–B16×(SIN(E26)))/B15)^2
D11×H9+E11×H10+F11×H11	= z' =	0	(–)Sin(–15×p/180)	Cos(–15×p/180)		0	= z''			
								X	Y	Az. angle
Matrix multiply			Pitch = –5					(–K10+SQRT(K10^2–4×J10×L10))/(2×J10)	(D28/D29–B14×J48–B16×(SIN(E26)))/B15	180/PI()×ATAN(K13/J13)
D14×H14+E14×H15+F14×H16	= x =	Cos(–5×π/180)	0	Sin(–5×π/180)		x'				ATAN(Y/X)
D15×H14+E15×H15+F15×H16	= y =	0	1	0	×	y'				
D16×H14+E16×H15+F16×H16	= z =	(–)Sin(–5×π/180)	0	Cos(–5×π/180)		z'				
								Roll, pitch, raw:		
Matrix multiply			Yaw =–3							
D19×H19+E19×H20+F19×H21	= xyaw =	Cos(–3×π/180)	Sin(–3×π/180)	0		x		A	B	C
D20×H19+E20×H20+F20×H21	= yyaw=	(–)Sin(–3×π/180)	Cos(–3×π/180)	0	×	y		1+(B19/B20)^2	(–)2×B19/B20^2×(D28/D29—B21×SIN(E26))	(SIN(E26))^2–1+((F29–B21×SIN(E26))/B20)^2
D21×H19+E21×H20+F21×H21	= zyaw=	0	0	1		z				
								X	Y	Az. angle
								(–K20+SQRT(K20^2-4×J20×L20))/(2×J20)	(D28/D29–B19×J23–B21×SIN(E26))/B20	180/PI()×ATAN(K23/J23)
								Note that the Yaw is simply an offset of 3°		
			Radians	**Azimuth:**				**Error:**		
Elevation angle (deg.) =		45	0.785398163	24				= F26 – J28		
Electrical phase diff. (deg.) =		635.295	= D27/D29	Elect. phase/phase gain				**Answer:**		
Electrical phase diff. (rad) =		11.08798947	38.60286059					= L23		

15.8 Alternate method

The previous analysis was done with the x-axis on the bow–stern axis and the y-axis on the port–starboard axis. If another axis is used, then the equations need to be modified to reflect the correct axis. If the x-axis is on the port–starboard axis and the y-axis on the bow–stern, then a right-hand coordinate system is defined. This analysis can be used with the following changes:

1. The quadratic equation for solving x needs to be:

$$X = \frac{-B - \sqrt{B^2 - 4AC}}{2A}$$

2. The azimuth angle is calculated as:

$$\varphi = a\tan(X/Y)$$

Note: The formula for converting the differential phase in electrical degrees to angular degrees is:

$$\text{Angular\_degrees} = (1/\text{phase\_gain}) \times \text{differential\_phase}$$

where angular_degrees is the number of degrees in azimuth that the target is from the boresight and right of boresight is defined as positive, phase_gain is the $(2\pi d)$/wavelength, and differential_phase is the difference in phase between the two antennas.

This parameter is measured in electrical degrees.

15.9 Quaternions

Quaternions are often used in place of coordinate conversions. The advantages of using quaternions are to prevent ambiguities in the coordinate conversion process and possibly to make it more efficient to implement. The coordinate conversion process compensates for the movement, roll, pitch, and yaw in three matrix multiplication around three axis. Quaternions compensate the roll, pitch, and yaw by rotating the position around one axis that is calculated in the process.

15.10 Summary

Interferometers use phase differencing to calculate the direction of the source of transmission. The basic mathematical tool used in interferometer calculations is the direction cosine. Direction cosines define the direction of the phasor in a three-dimensional plane. The basic interferometer equation deals with only two-dimensional analysis. The third dimension alters the two-dimensional solution significantly. This three-dimensional approach using coordinate conversion techniques to compensate for baseline rotations provides an accurate solution for the

three-dimensional interferometer which includes the elevation angle from the ground.

15.11 Problems

1. What is the true azimuth angle for a two-dimensional interferometer given that the operational frequency is 1 GHz, the phase difference between the two antennas is 10 radians, and the antennas are separated by 3 m?
2. Why is a three-dimensional approach required for a typical interferometer calculation?
3. Why is the $\sin\theta$ used for the elevation angle calculation?
4. Is the order of the coordinate conversion process critical? Why?

Further reading

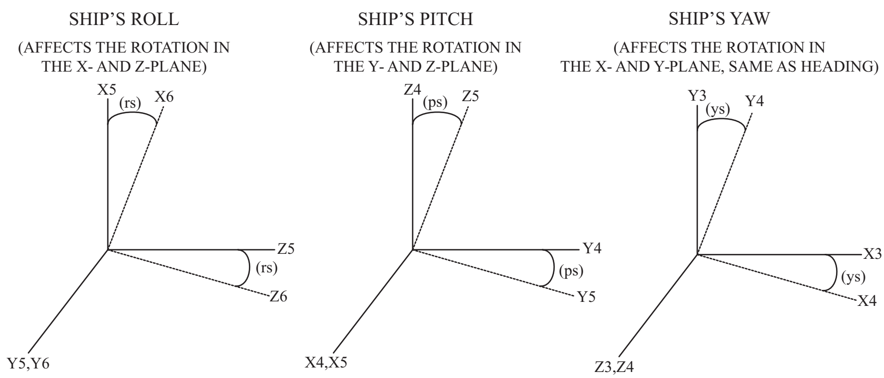
Halliday, David, Robert Resnick, and Jearl Walker. *Fundamental of Physics*, 9th ed. Hoboken, NJ: John Wiley & Sons, 2010.

Salas, Saturnino L., and Einar Hille. *Calculus: One of Several Variables*, 10th ed. Hoboken, NJ: John Wiley & Sons, 2007.

Shea, Don. "Notes on Interferometer Analysis." 1993.

Appendix A
Coordinate conversions

The following are diagrams showing the coordinate conversions for motion including heading, roll, pitch, and yaw. These conversions are for the alternate method using the *x*-axis on the port–starboard plane and the *y*-axis on the bow–stern plane (Figure A.1).



$$\begin{pmatrix} X6 \\ Y6 \\ Z6 \end{pmatrix} = \begin{pmatrix} \cos(rs) & 0 & \sin(rs) \\ 0 & 1 & 0 \\ -\sin(rs) & 0 & \cos(rs) \end{pmatrix} \begin{pmatrix} X5 \\ Y5 \\ Z5 \end{pmatrix}$$

$$\begin{pmatrix} X5 \\ Y5 \\ Z5 \end{pmatrix} = \begin{pmatrix} \cos(rs) & 0 & -\sin(rs) \\ 0 & 1 & 0 \\ \sin(rs) & 0 & \cos(rs) \end{pmatrix} \begin{pmatrix} X6 \\ Y6 \\ Z6 \end{pmatrix}$$

$$\begin{pmatrix} X5 \\ Y5 \\ Z5 \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 & 0 \\ 1 & \cos(ps) & 0 & -\sin(ps) \\ 0 & \sin(ps) & 0 & \cos(ps) \end{pmatrix} \begin{pmatrix} X4 \\ Y4 \\ Z4 \end{pmatrix}$$

$$\begin{pmatrix} X4 \\ Y4 \\ Z4 \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 & 0 \\ 1 & \cos(ps) & 0 & \sin(ps) \\ 0 & -\sin(ps) & 0 & \cos(ps) \end{pmatrix} \begin{pmatrix} X5 \\ Y5 \\ Z5 \end{pmatrix}$$

$$\begin{pmatrix} X4 \\ Y4 \\ Z4 \end{pmatrix} = \begin{pmatrix} \cos(ps) & -\sin(ps) & 0 \\ \sin(ps) & \cos(ps) & 0 \\ 0 & 0 & 1 \end{pmatrix} \begin{pmatrix} X3 \\ Y3 \\ Z3 \end{pmatrix}$$

$$\begin{pmatrix} X3 \\ Y3 \\ Z3 \end{pmatrix} = \begin{pmatrix} \cos(ps) & \sin(ps) & 0 \\ -\sin(ps) & \cos(ps) & 0 \\ 0 & 0 & 1 \end{pmatrix} \begin{pmatrix} X4 \\ Y4 \\ Z4 \end{pmatrix}$$

Figure A.1 Coordinate conversions

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Appendix B

True north calculations

B.1 True north calculations

The formula to be used for correcting from true north is

$$\text{Bearing\_true}(n) = \text{Relative\_bearing}(n) + \text{ships\_heading}(n)$$

where bearing_true(n) is the target bearing from the ship with respect to true north, ships_heading is the ship's heading as measured by the ship's inertial navigation system or a magnetic compass, and Relative_bearing(n) is the current target bearing relative to ship's longitudinal axis starting at the bow for zero degrees relative and rotating clockwise.

The formula to be used for correcting from true north to magnetic north is

$$\text{Bearing\_mag}(n) = \text{bearing\_true}(n) + \text{magnetic\_variation}$$

where bearing_mag(n) is the target bearing from the ship with respect to magnetic north, bearing_true(n) is the target bearing from the ship with respect to true north, and magnetic_variation is the variation between magnetic north and true north. Magnetic variation may be plus or minus.

B.2 Phase ambiguities

For an interferometer to have no phase ambiguities, the spacing between the antennas should be less than $\lambda/(2\pi)$ wavelength apart. This provides a phase gain of less than 1. Note that a phase gain of exactly 1 gives a one-to-one conversion from azimuth angle to electrical angle. Therefore, there are no phase ambiguities. If the separation is greater than 1 wavelength, giving a greater than 1 phase gain, then ambiguities exist. For example, if the separation is 2λ, then half the circle covers 360° and then repeats for the second half of the circle. Therefore, a phase measurement of 40° could be two spatial positions.

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Appendix C
Elevation effects on azimuth error

C.1 Elevation effects on azimuth error

The elevation effects on the azimuth error are geometric in nature and were evaluated to determine if they are needed in the azimuth determination and to calculate the magnitude and root-mean-square azimuth error. A simulation was done using three different angles 10°, 25°, and 45°. The simulation sweeps from 0° to 24° in azimuth angle, and the error is plotted in degrees. The results of the simulations are shown in Figure C.1.

The azimuth error is directly proportional to the elevation angle; the higher the angle, the greater the error. Also, the azimuth error is directly proportional to the azimuth angle off interferometer boresight. This analysis was performed with a horizontal interferometer baseline with no pitch and roll. The azimuth error at an azimuth angle of 22.5° and an elevation angle of 45° was equal to 6.8°, which is the worst case error.

The azimuth error caused by elevation angle can be calculated by

$$\begin{aligned}\text{Az Error} &= \text{True Az} - \text{asin}\left[(\cos(\text{true el})\sin(\text{true az}))/\cos(\text{assumed El angle})\right]\\ &= 22.5 - \text{asin}[(\cos(45)\sin(22.5))/\cos(0)] = 6.8^\circ\end{aligned}$$

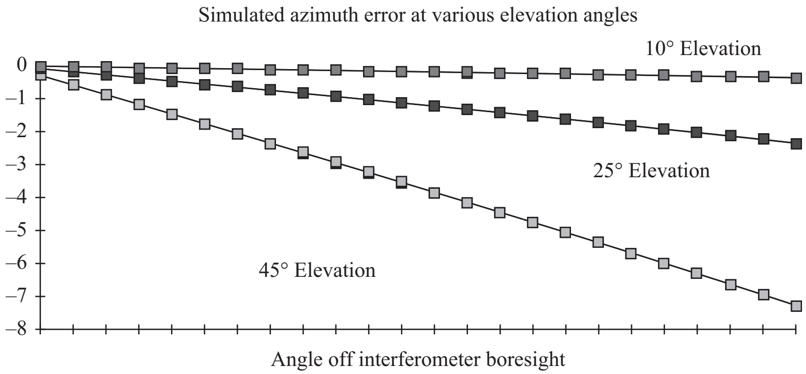

Figure C.1 Azimuth error due to elevation angle

For elevation compensation only, an approximate solution can be used:

$$\mathrm{dp} = (2\pi d)/\lambda \sin\theta \cos\Psi$$

where Ψ is the elevation angle.

However, including roll and pitch produces a three-dimensional analysis, using the standard interferometer, two-dimensional equation, tries to compensate for the elevation, roll, and pitch, which results in a very complex transcendental equation, without making too many assumptions, and is generally solved by iterative methods.

Appendix D

Earth's radius compensation for elevation angle calculation

D.1 Earth's radius compensation for elevation angle calculation

The angle (θ), which is the desired elevation angle, is calculated for the curved earth as shown below:

Solve for the angle α using the law of cosines:

$$c^2 = a^2 + b^2 - 2ab\cos(\alpha)$$
$$\alpha = a\cos((a^2 + b^2 - c^2)/2ab)$$

where a is the slant range, b is the altitude of the ship plus the earth's radius, and c is the altitude of the aircraft plus the earth's radius (Figure D.1).

Therefore,

$$\theta = \alpha - 90^\circ = a\cos((a^2 + b^2 - c^2)/2ab) - 90^\circ$$

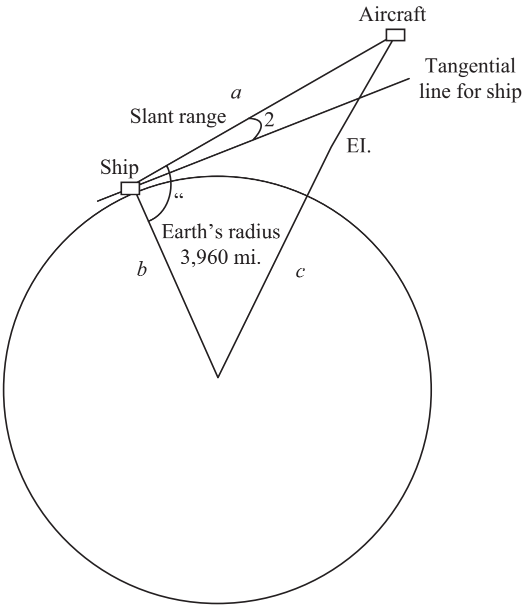

Figure D.1 Earth's radius compensation for elevation angle calculation

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Answers

Chapter 1

1. Answer: 0 dBm is a power level of 1 mW. 2 dBm is a power level of 1.58 mW. 0 dBm ± 2 dBm = 1 mW ± 1.58 mW = 1 mW + 1.58 mW = 2.58 mW = 4.1 dBm and 1 mW − 1.58 mW = −0.58 mW can't have negative power.
   The correct expression is 0 dBm ± 2 dB = ±2 dBm = +2 dBm = 1.58 mW and −2 dBm = 0.63 mW.
2. Answer: Bit rate = Bandwidth.
3. Answer: 1 nW/3.16 × 316/6.3 × 31.6/2 = 0.00025 mW. 10 log(0.00025) = −36 dBm. Or: −60 dBm − 5 dB + 25 dB − 8 dB + 15 dB −3 dB = −36 dBm.

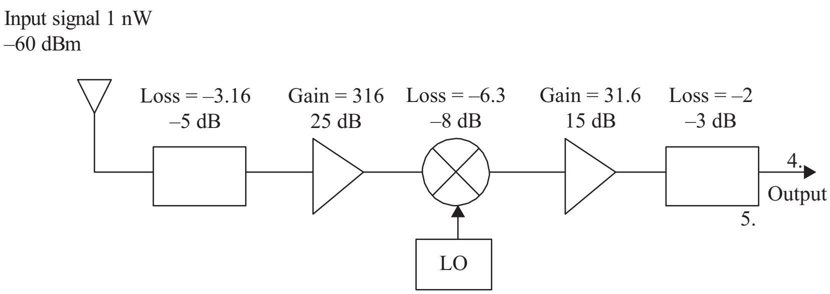

Figure 1.P2 Receiver diagram

4. Answer: 10 log 1 mW = 0 dBm. 10 log(0.01 mW) = −20 dBm. Therefore, the spurious response is −20 dBm − 0 dBm = −20 dBc. For −40 dBc, the spur level is −40 dBc + 0 dBm = −40 dBm = 0.1 μW.
5. Answer: The MDS will be reduced by 1.5 dB. Increase the transmitter power by 1.5 dB, reduce the losses after the power amplifier of the transmitter and before the LNA of the receiver by 1.5 dB or reduce the noise figure of the LNA by 1.5 dB.
6. Answer: $G_t = 30$ dBi $= 10 \log[n(\pi D/\lambda)^2]$. $1{,}000 = n(\pi D/\lambda)^2$. Therefore, $D = (1{,}000/n)^{1/2} \times \lambda/\pi = (1{,}000/0.5)^{1/2} \times 0.06\ \text{m}/\pi = 0.854$ m.
7. Answer: 125.4 dB of attenuation. $A_{fs} = 20 \log[4\pi R/\lambda]$, 10 nmi = 18,520 m. $A_{fs} = 20 \log(4\pi(18{,}520)/0.125) = 125.4$ dB attenuation.
8. $v_r = 27.8$ m/s, lambda = 0.3/2.4 = 0.125, $f_{\text{Doppler}} = 27.8/0.125 = 222$ Hz.

9. $v_r = 10$ m/s $\times \cos(45°) = 7.07$ m/s, lambda $= 0.3/5.8 = 0.0517$, $f_{\text{Doppler}} = 7.07/0.057 = 124$ Hz.
10. Answer: $\text{KT} + 10\ \log(BW) + \text{NF} = -174\ \text{dBm} + 10\ \log(10\ \text{MHz}) + 3\ \text{dB} = -101\ \text{dBm}$.
11. Answer: You must convert to actual power and use the noise factor equation and then convert the answer to dB. The results are as follows:

$$\begin{aligned} F_t &= F_1 + [(F_2 \times \text{Losses}) + -1]/G_1 \\ &= 1.995 + [(3.98 \times 3.16 - 1)/100] = 2.11: \\ \text{Noise Figure} &= 10\log(2.11) = 3.24\ \text{dB}. \end{aligned}$$

12. Connects the transmitter to the receiver and determines if there is enough S/N to complete the link, also used for trade-offs.

Chapter 2

1. Crest Factor is the $P_{\text{peak}}/P_{\text{rms}}$. Takes higher power to transmit the P_{peak}.
2. Load impedance = source impedance.
3. Perfect reconstruction of the digital waveform.
4. Software programable radio. Monitors the environment and uses an SDR to change parameters.
5. Binary Phase Shift Keying 0, 180. 16 Quadrature Amplitude Modulation, uses quadrature phase with 2 levels of amplitude. BPSK—robust, 16 QAM high data rates.
6. Error Voltage Magnitude. Measures how far the noise distorts the desired signal.
7. No sidelobe regeneration, spectrally efficient, operates in saturation.
8. Sinusoidal weighted OQPSK, Minimum separated FSK.
9. Gaussian, raised cosine, raised cosine squared, root raised cosine.
10. Uses more spectrum than is needed to send the data.
11. Antijam, LPI/LPD, multipath. More loss due to spreading losses.
12. Phase shift keying, Frequency Hopping.
13. Ratio of the PN-code bandwidth and the data bandwidth.
14. Tapped delay line and an XOR.
15. Repeating codes.
16. Randomly invert the codes.
17. TDMA, FDMA, CDMA.
18. To keep the codes separated in the receiver.
19. Differential uses change between last state and present state. Reduces the effects of instability in oscillators, phase noise. Coherent has better BER.
20. Orthogonal Frequency Division Multiplexing. Allows overlap between users.
21. Eliminates the near-far problem.
22. Required Eb/No and Spreading losses.
23. High-speed code modulo-2 with the data.

Chapter 3

1. Contains a double conversion, two stage converters.
2. A circulator or a T/R switch.
3. $G_r = 20\ \log[V_o/2^n] - \text{MDSI} = 20\ \log[1/2^8] - [-114 - 10\ \log 10 + 3 + 4] =$ 68.8 dB.
4. Third-order SFDR $= 2/3(\text{IP}_3 + 174 - \text{NF} - 10\ \log B) = 2/3(20 + 174 - 3 - 10\ \log(10\ \text{MHz})) = 80.66$ dB.
5. $80.66 - 10 = 70.66$ dB.
6. 3 dB + 10 dB = 13 dB.
7. $1 \times 0 = 10$ MHz
   $0 \times 1 = 12$ MHz
   $2 \times 0 = 20$ MHz
   $0 \times 2 = 24$ MHz
   $3 \times 0 = 30$ MHz
   $0 \times 3 = 36$ MHz
   $1 \times 1 = 2$ MHz, 22 MHz
   $1 \times 2 = 14$ MHz, 34 MHz
   $2 \times 1 = 8$ MHz, 32 MHz.
8. 120 MHz/100 MHz = 1.2 MHz.
9. According to the Nyquist criteria, Sample rate $= 2 \times 1$ MHz = 2 Msps.
10. Maximum phase error $= 45 - \tan^{-1}(1/2) = 18.43°$
    Maximum amplitude error = 6 dB.
11. The advantages for oversampling the received signal are better resolution for determining the signal and better accuracy for determining the time of arrival. The disadvantages are generally more expensive and require more processing for the increased number of samples taken.
12. So that the quality of the time domain square pulse signal is preserved. The result of nonconstant group delay is distortion or dispersion of the pulse which causes intersymbol interference (ISI).
13. LNA stands for low-noise amplifier. It is important because it is the main contributor to the noise figure of the receiver and is a direct correlation to the link budget. 1 dB higher noise figure results in 1-dB lower link margin.
14. Aliasing occurs where the high frequencies can alias or appear as lower frequencies and distort the design signal.
15. Each ADC bit divides the voltage range by half, and half the voltage is equivalent to 6 dB.

Chapter 4

1. Voltage-controlled attenuator and a voltage-controlled amplifier.
2. The RC time constant should be much larger than the period of the carrier and much smaller than the period of any desired modulating signal. The period of the carrier is 0.1 µs, 1/10 MHz, and the desired modulating signal period is

1 μs, 1/1 MHz. Therefore, the period should be about 0.5 μs, 2 MHz. This will depend on how much distortion of the desired signal there is compared to how much of the carrier is allowed through the processing.

3. The integrator provides a zero steady-state error for a step response. Constant output level to the detector with no variation.
4. As the error approaches steady state (very slow-changing error), then the gain of the integrator approaches infinity. Therefore, any small change in error will be amplified by a very large gain and drives the error to zero.
5. An integrator.
6. If the slope is nonlinear, using a linear approximation causes the loop gain in some operational periods to be less or more than the approximation; therefore, the response time will be slower or faster, respectively.
7. The diode is nonlinear and is the point where the piecewise linear connections are. It provides a smoothing function due to the nonlinearity.
8. The integrator for the PLL is built in and does not need to be added, whereas the AGC needs to have an integrator added to the circuit.
9. They both are feedback systems. They just have different parameters that are in the feedback loop.
10. The PLL is in the lock state. This analysis does not include the capture state where the PLL has to search across a wider bandwidth to bring it into the lock state.
11. In all feedback systems, careful design needs to be done to prevent oscillations or instability.

Chapter 5

1. A squaring loop and a Costas loop.
2. $[\mathrm{Cos}(\omega t + (0,\pi))]^2 = \mathrm{Cos}^2(\omega t + (0,\pi)) = 1/2[1 + \mathrm{Cos}(2\omega t + 2(0,\pi))] = 1/2[1 + \mathrm{Cos}(2\omega t + (0,2\pi))] = 1/2[1 + \mathrm{Cos}(2\omega t + 0)]$. Therefore, the phase ambiguity is eliminated. However, the frequency needs to be divided in half to obtain the correct frequency.
3. (a) Signal would need to be squared three times. (b) Signal would need to be squared four times.
4. Decision directed Costas loop.
5. A matched filter correlator and sliding correlator.
6. The incoming code comes into the correlator and when the spread spectrum codes line up, a large pulse is generated with the spread spectrum removed.
7. $2\times 1/\mathrm{PW} = 2\times 1/1\ \mu\mathrm{s} = 2$ MHz.
8. Pulse Position Modulation. Coherent is hard to keep locked over a long period of time but it has less errors. Differential is much easy, not affected by phase delays, Doppler, and poor oscillators.
9. The receiver slides the stored code in time until it lines up with the incoming code. Used to eliminate the Spread spectrum code.
10. An early-late gate.

11. BW for match filter correlator is the same, sliding correlator reduces BW.
12. Retraces of the signal on the oscilloscope with a PN sequence. Band limiting makes it look like an eye.
13. The best place is the center of the eye. The worst place is transition point of the eye.
14. All frequencies have the same delay. Causes dispersion and ISI and interferes with detection.
15. Intersymbol interference is the distortion caused by a chip or bit that interferes with the other chips or bits, usually with the adjacent chips or bits.
16. PN code that scrambles the waveform so there are not long runs of 0 or 1 s.
17. SNR and bandwidth.
18. Squaring twice, same as with QPSK.

Chapter 6

1. Since the integral of the probability density function is equal to $1 = 100\% = P_{\text{wrong}} + 37\%$; $P_{\text{wrong}} = 63\%$.
2. $E[x] = \Sigma\, xf_x(x) = 1(0.4) + 2(0.6) = 1.6$. The mean is equal to $E[x] = 1.6$.
3. Closer to 2 because there is a higher probability that the answer is going to be 2 than 1.
4. $E[x^2] = \Sigma\, x^2 f_x(x) = 1(0.4) + 4(0.6) = 2.8$.
5. $\text{Var} = E[x^2] - \text{mean}^2 = 2.8 - 2.56 = 0.24$.
6. $\text{Std Dev} = (\text{var})^{1/2} = (0.24)^{1/2} = 0.49$.
7. $1 - 0.954 = 0.046 = 4.6\%$.
8. Increase the number of ADC bits.
9. The probability receiving 1 pulse is 0.98. The probability of receiving all 20 pulses is

$$0.98^{20} = 66.8\%.$$

10. Using the binomial distribution function the probability of only one error is

$$p(19) = \binom{20}{19} p^{19}(1-p)^{(20-19)} = \binom{20}{19}(0.98)^{19}(0.02)^1$$
$$= \frac{20!}{(20-19)!19!}(0.98)^{19}(0.02)^1 = 26.7\%$$

Therefore, the percentage of the errors that result in only 1 pulse lost out of 20 is

$$\text{Percent(1pulse lost)} = 66.8\%/(100\% - 26.7\%) = 48.9\%.$$

11. Percent (1 pulse lost) = 66.8%/(100% − 26.7%) = 48.9%. The probability of error is a predicted value of what you should get given an Eb/No value, Bit error rate is the actual measurement of the system, number of errors of the total number of bits sent.

12. Parity, check sum, cyclic redundancy check CRC. CRC is the best.
13. Block codes and convolutional codes.
14. Since interference causes burst errors, interleaving disperses these errors over multiple messages so that fewer errors need to be corrected for a given message.

Chapter 7

1. Glint errors are angle of arrival errors and scintillation errors are amplitude fluctuations.
2. Specular multipath affects the solution the most since it is more coherent with the signal and directly changes the amplitude and phase. Diffuse multipath is a constantly changing more random signal which looks more like noise and is usually smaller in amplitude.
3. The effect of multipath at the pseudo-Brewster angle for vertically polarized signal is reduced.
4. There is basically no effect of multipath at the pseudo-Brewster angle for horizontally polarized signal. Therefore, the multipath still highly affects the incoming signal.
5. Rayleigh criterion.
6. The Rayleigh criterion is as follows:

$$h_d \sin\, d \; < \frac{1}{8}$$

where

h_d is the peak variation in the height of the surface and d is the grazing angle.

If the Rayleigh criterion is met, then the multipath is a specular reflection on a rough surface. Therefore,

$$\begin{aligned} 10\,\sin(10) &= \frac{\lambda}{8} \\ \lambda &= 80\,\sin(10) = 13.89\text{ m} \\ f &= \frac{c}{\lambda} = \frac{3\times 10^8}{13.89} = 21.6\text{ MHZ} \end{aligned}$$

7. The divergence factor is the spreading factor caused by the curvature of the earth. This factor spreads out the reflecting surface. This is generally assumed to be unity since the effects are negligible for most applications, except for possibly satellites.
8. Leading edge tracking. The multipath returns are delayed from the desired return, so the radar detects the leading edge of the pulse and disregards the rest of the returns.

9. The vector addition is shown in the figure:

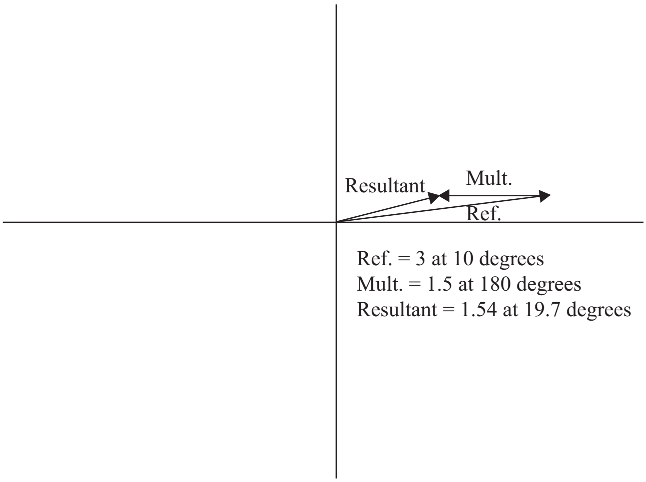

Figure 7.P9 Vector addition showing multipath effects

10. The power summation uses power instead of voltage. The reflection coefficient is squared to represent a power reflection coefficient.
11. Since multipath is spatial dependent, antenna diversity uses multiple antennas at different spatial positions. If one antenna is in a multipath null, there is a high probability that the other antennas are not and the signal is reliably received.

Chapter 8

1. A pulse jammer duty cycle is equal to 1/response time of the AGC. Therefore, with a jammer present, the AGC adjusts the gain to minimum, then the jammer turns off and it takes the response time of the AGC to recover, and then the jammer turns on again. This captures the AGC and optimizes the jammer.
2. Approximately 1 MHz.
3. The FIR filter has fixed weight values for a given filter response, and the adaptive filter uses feedback to update or adjust the weight values.
4. The unwanted sidebands during the mix down and up need to be eliminated.
5. The μ value is the gain of the feedback process.
6. Increasing the μ value does the following:
   (i) Increases convergence time.
   (ii) Generally decreases stability.
   (iii) The steady-state error is larger.

7. The assumption is good when the jammer is much larger than the signal, and the directional antenna has significant gain toward the desired signal and provides a reduction in J/S.
8. The assumption is bad when the jammer is not larger than the signal and the directional antenna does not provide a reduction in J/S.
9. Channelized receiver. This receiver covers the largest instantaneous bandwidth with the best sensitivity.
10. Adaptive filter.

Chapter 9

1. Monitor and control with learning and reasoning.
2. Power, frequency, modulation.
3. SDRs.
4. Dynamic spectrum access or dynamic spectrum allocation. Base monitors the spectrum and changes the frequency to prevent jamming. Remote follows the base.
5. Both the base and remote switch to an *a priori* frequency in a random sequence.
6. Changes power to required threshold set in the base. Power control contains in a large feedback loop which includes the remote.
7. Closed loop uses feedback with RSS. Open uses Nav or other data.
8. Base.
9. Delta power received.
10. Adds a gain factor in the loop.
11. Base controls the power for each of the remotes.
12. Data rate and robustness against interference.
13. Null Steering, Antenna steering, etc.
14. Increase data rate or improved robustness to jammers.
15. Multihop.
16. Desired signal uses multipath as the primary channel.
17. Mobile Ad hoc Network.
18. Physical, cognitive radio.
19. Network uses multihop around the interference.
20. Learning, predicting.
21. Game theory.
22. If all the strategies are known, any user cannot benefit by changing their strategy without affecting the network or other users.
23. Monitor & Control. Monitors the environment and controls the capabilities to improve performance.
24. Detection, perform trade-offs, optimize solution, device selection, change and control devices, reason and learn. It uses all the capabilities and implements the best or least cost solution.
25. Regulations and rules to prevent interference on dynamic systems.

Chapter 10

1. Raster, binary, random scan.
2. Random bias.
3. Outside-in approach.
4. Inside-out approach.
5. Beam spoiling, use of the sidelobes.
6. Eb/No is reduced by 6 dB. Generally yes, depends on the required Eb/No that is needed to detect the signal which is much less than is required to maintain the probability of error.
7. Sequential scanning.
8. Sequential lobing.
9. Provides a conical scan around boresite, and the closer the scan is to the user the larger the return which tells the antenna where to point.
10. At boresite.
11. Antenna contains 4 quadrants that determine which quadrant has the largest signal and test the antenna to point in that direction.
12. Same as answer 11 with the beam confined to just the main lobe
13. α/β Tracker.
14. Open-loop tracker provides the position with no feedback, does not depend on the previous position. Closed-loop tracker relies on the previous position and uses feedback to update the position. Select α and β values with expected dynamics.
15. Weighted or bias integration of the two tracks.
16. Closed loop or alpha-beta bias or weighting for long range, Open-loop or Navigational bias or weighting for short range.

Chapter 11

1. Power line, phone line, and RF.
2. Orthogonal frequency division multiplexing (OFDM).
3. The inner product is equal to zero.
4. The inner product of the signal with itself is equal to one. For orthogonal signals, the inner product is zero. Therefore, taking the inner product with a duplicate of the known signal will produce only the desired signal.
5. A phone line needs to be present and not all homes have phone lines going to every room in the house.
6. Interoperability, antijam, and security.
7. SDRs can be programed on the fly to produce different waveforms and modulation schemes so they can be programed for different users.
8. Star, bus, ring, and mesh.

Chapter 12

1. Satellite system is the most ubiquitous infrastructure since it covers the globe.
2. Geosynchronous satellites are synchronous with the earth's rotation, and geostationary satellites are geosynchronous satellites with their orbits on the equatorial plane. Therefore, they appear as if they are stationary with respect to the earth. Geosynchronous satellites usually have inclined orbits so they move around with respect to the earth.
3. The figure-eight pattern around the orbital plane that increases with age of the satellites. This phenomenon allowed a point on the South Pole to see the satellites for a period of time at the bottom of the figure-eight pattern for communications.

Chapter 13

1. C/A code = 2 × 1.023 MHz = 2.046 MHz
   P-code = 2 × 10.23 MHz = 20.46 MHz.
2. Theoretical process gain for C/A code using the data rate of 50 Hz = 10 log 1.023 MHz/50 = 43 dB.
   Theoretical process gain for P code using the data rate of 50 Hz = 10 log 10.23 MHz/50 = 53 dB.
3. Short-length code for faster acquisition times. Slower rate code for a higher signal to noise in a smaller bandwidth.
4. Long-length code providing a more covert signal for detection. Faster rate code for higher process gain against unwanted signals.
5. The narrow correlator is better accuracy of the GPS solution.
6. The narrow correlator is less stable, is easier to jam due to the wider bandwidth required to process the signal, and provides no benefit using a P-code receiver due to the fact that the bandwidth is already limited in the transmitter.
7. Jitters the clock timing and distorts ephemeris data regarding the orbits of the SVs.
8. Carrier smoothing is using the carrier change of phase with time to filter the code measured range data with time.
   Carrier data is not affected by the SA, multipath, and ionospheric effects as much as the code is affected.
9. Common errors, for example, ionospheric errors, exist in both receivers and are subtracted out in the solution.
10. Wavelength ambiguity and cycle slips.
11. The wavelength of the difference frequency is larger; therefore, there are fewer wavelength ambiguities to search over.
12. Use one code and exclusive-OR with different delayed versions of that same code. This generates multiple codes dependent on the delay.
13. Differential and relative GPS techniques.

Chapter 14

1. RAdio Detecting And Ranging (RADAR), WWII.
2. Pulsed, continuous wave.
3. Pulse Radar: Advantages: Single Antenna, lower average power, range and altitude. Disadvantages: susceptible to jammers.
4. Short PRI causes distortion of the echo returns after the next pulse.
5. Pulse repetition interval; PRI = 1/PRF.
6. Duty Cycle = PW/PRI $\times$ 100 = PW $\times$ PRF $\times$ 100 = 1 μs $\times$ 100 kHz $\times$ 100 = 10%
7. Spectrum—fundamental in main lobe, even harmonics suppressed, odd harmonics centered in side lobes.
8. On/off keying, pulsed 100% AM.
9. 5.38 s/2 $\times$ 1125 ft/s = 5288 ft = 1 mi.
10. Radar crosssection.
11. Travels over and back through the same atmosphere.
12. 60 dBm + 40 dB + 40 dB − 100 dB − 100 dBm + 21 dB − 10 dB = −49 dBm.
13. Four times.
14. Speed of light.
15. $R_{\min} = (1\ \mu\text{s} + 0) \times 3 \times 10^8/2 = 150$ m.
16. Echoes from the first pulse comes after the second pulse causing errors in range.
17. Pulse width.
18. Ability to separate or resolve two equal targets, same bearing, different ranges.
19. Reduce the pulse width. Use pulse compression—dependent on the BW.
20. Angle of arrival to the target.
21. Linear velocity is a straight line, angular velocity follows the circumference of the circle.
22. Pulse position indicator. The target's range and bearing on a circular display.
23. Displays the amplitude of the echo where the range gate is placed.
24. Radar transmits pulses and receives target echoes to determine range and bearing.
25. Active electronic scanned array. Very fast beam and null steering, spoiling/narrowing, multiple beams, MIMO, no mechanically moving parts.
26. Monostatic uses same Tx&Rx antennas or colocated antennas. Bistatic uses separate antennas separated by a distance equal to the expected target range.
27. Uses two pulses at two different frequencies. Advantages: Sums both signals—3 dB of gain, increase range, smoothens fluctuations, increases probability of detection, max backscatter at one freq, min backscatter of other freq. Disadvantages: Higher probability of unwanted detection, uses more freq and bandwidth.

28. Uses one antenna with four compartments and takes the sum and differences to determine position on one pulse.
29. No scanning required. Completes in one pulse. Less susceptible to jamming.
30. Chirp radar. Sounds like a bird chip at audio frequencies. Acoustic waves. Lower frequencies are easier to control the delays. Transmitted signal is up-chirp, jammer is not. Receiver uses down-chirp, signal is retrieved, jammer is spread by the down-chip. Narrow band filter retrieves the signal.
31. Unwanted RF returns. Use MTI—Doppler processing, polarization.
32. P_{fa} = Probability the signal is above the detection threshold that is not a valid return. P_d = Probability the valid echoes are detected. Between the P_{fa} and P_d.
33. Moving target indicator. Doppler.
34. Moving targets that are cancelled the same as still targets. Occurs when the change in range between pulses is exactly one-half of the wavelength, changing the phase by 360° which is the same as 0°, or no phase shift at all. Appears the moving target is stationary.
35. Multiple pulse radar using staggered pulses and irregular PRIs.
36. CP-PSK. Use in saturation without sidelobe regeneration.
37. Weather.
38. Synthetic-Aperture Radar. Resolution Imagery. For aircraft and spacecraft.
39. $P_t = 1$ W = 30 dBm, frequency = 2.4 GHz, $\lambda = 3 \times 10^8/2.4 \times 10^9 = 0.125$, $R = 100$ m, calculate freespace loss in dBm. $A_{fs} = 20 \log \lambda/(4\pi R) = 20 \log 0.125/(4\pi 100) = -80$ dB: $P_r = 30$ dBm + (−80 dB) = −50 dBm (Add), $A_{\text{fs}} = 20 \log (4\pi \text{R})/\pi = 20 \log (4\pi 100)/0.125 = 80$ dB: $P_r = 30$ dBm − (80 dB) = −50 dBm (Subtract).
40. PRF = 1/PRI = 1/2 μs = 500 kHz, 5 cycles = 1 PW = 100 ns. 1 cycle = 100 ns/5 = 20 ns. f_o = 1/20 ns = 50 MHz, duty cycle = PW/PRI = (100 ns/2 μs) × 100 = 5%.
41. $P_r = 100 \times (316)^2(0.125)^2 \times 110/[(4\pi)^3 \times 100^4 \times 10] = 8.65\times10^{-6}$ mW.
42. $A_{fs\text{dB}} = 20 \log[(0.125)/(4\pi 100)] = -80$ dB, $G_{\text{targ}} = 10 \log(4\pi\times 110/0.125^2) = 49.5$ dB, $P_{r\text{dBm}} = 20$ dBm + 2 × 25 dB + 2× (−80 dB) + 49.5 dB − 10 dB = −50.5 dBm.
43. $R_{\max} = (\text{PRI} - \text{T}) \times c_0/2$, $R_{\max} = (15\ \mu\text{s} - 100\ \text{ns}) \times 3 \times 10^8/2 = 2.235$ km.
44. $S_r = (c_0 \times \tau)/2 = c_0/2B_{\text{tx}} = 3 \times 10^8/(2 \times 1\ \text{MHz}) = 150$ m.
45. $S_A = 2 \times R \times \sin(\theta/2)$ m $= 2 \times 100 \times \sin(3/2) = 5.2$ m.
46. $\theta = \sin^{-1}(\lambda/D) = \sin^{-1}(0.125/3) = 2.39°$.
47. $\omega = v/r = 10/20 = 0.5$ rad/s, $\omega = 0.5$ rad/s $\times 180/\pi = 28.6°$/s.
48. 70 km/h × 1 h/3,600 s = 19.4 m/s, $f_{\text{Doppler}} = 2\text{v}_\text{r}/\lambda_0 = 2\times 19.4/0.125 = 310$ Hz.
49. $v_r = v \times \text{Cos}(\theta) = 10$ m/s × Cos(30°) = 8.66 m/s.
50. $v_b = 1\times\lambda \times \text{PRF}/2 = 0.125 \times 100/2 = 6.25$ m/s.
51. Doppler frequency $= 2v_r/c_0 \times f_{\text{transmit}} = (2\times 75/3\times 10^8) \times 1 \times 10^9 = 500$ Hz, Nyquist Criteria = 2 × (500 Hz) = 1 kHz < 2 kHz sampling rate, exceeds Nyquist criteria − Tracks Velocity at 75 m/s.

52. Doppler frequency = $2v_r/c_0 \times f_{\text{transmit}} = (2 \times 500/3 \times 10^8) \times 1 \times 10^9 = 3.3$ kHz, Nyquist Criteria = 2× (3.3 kHz) = 6.6 kHz < 2 kHz sampling rate, does not meet Nyquist criteria – does not track velocity at 500 m/s.
53. (a) Doppler frequency = $2v_r/c_0 \times f_{\text{transmit}} = (2 \times 160/3 \times 10^8) \times 1 \times 10^9 =$ 1.067 kHz, Nyquist Criteria = 2 × (1.067 kHz) = 2.13 kHz < 2 kHz sampling rate, does not track.
    (b) $v_r = v \times \cos(45) = 113$, Doppler frequency = $2v_r/c_0 \times f_{\text{transmit}} = (2 \times 113/3 \times 10^8) \times 1 \times 10^9 = 1.067$ kHz, Nyquist criteria = 2 × (753 Hz) = 1.507 kHz < 2 kHz sampling rate, tracks the target.

Chapter 15

1. $\lambda = 3 \times 10^8/1 \times 10^9 = 0.3$ m.

$$\theta = \text{asin}\,[dp/((2\pi d)/\lambda)] = \text{asin}\,[10/((2\pi 3)/0.3)] = 9.16^\circ.$$

2. The elevation angle affects the azimuth interferometer calculation.
3. The direction cosines are defined for the cos(a) from the top down. Therefore, the angle up is the sin θ, where θ is the elevation angle from the horizontal baseline.
4. The order is critical. There is a difference in the solution for the order of the coordinate conversion. For example, if the platform is pitched and then rolled, this gives a different solution from if the platform is rolled and then pitched.

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Transceiver and System Design for Digital Communications

5th Edition

This applied engineering reference covers a wide range of wireless communication design techniques; including link budgets, error detection and correction, adaptive and cognitive techniques, and system analysis of receivers and transmitters. Digital modulation and demodulation techniques using phase-shift keyed and frequency hopped spread spectrum systems are addressed. The book includes sections on broadband communications and home networking, satellite communications, global positioning systems (GPS), search, acquisition and track, and radar communications. Various techniques and designs are evaluated for modulating and sending digital signals, and the book offers an intuitive approach to probability plus jammer reduction methods using various adaptive processes. This title assists readers in gaining a firm understanding of the processes needed to effectively design wireless digital communication and cognitive systems with a basic understanding of radar.

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Derived from numerous training workshops taught to engineers through private courses by the authors, this book will appeal to digital wireless communications system designers in both the commercial and military sectors, in particular new engineers requiring practical design techniques and fundamental understanding of modern systems that employ digital transceivers.

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